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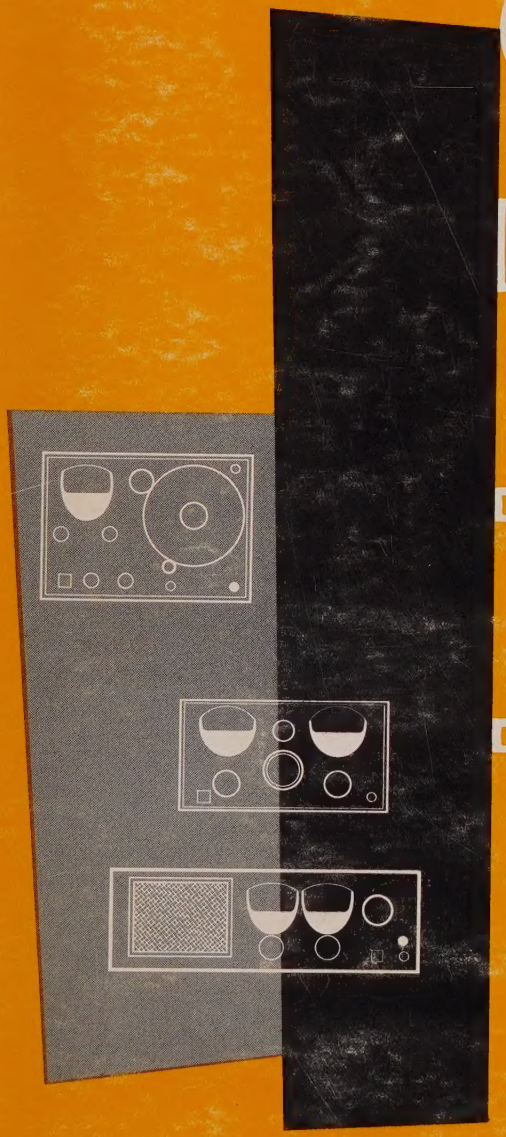
INSTRUCTION MANUAL

MODEL 315

RESISTOR-NOISE TEST SET

For Serial No._____



QT

- REGULATED POWER SUPPLIES
- ELECTRONIC INSTRUMENTS

Quan-Tech
LABORATORIES INC.
BOONTON, N. J.

INSTRUCTION MANUAL

MODEL 315

RESISTOR-NOISE TEST SET

For Serial No._____

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FACILITIES

Quan-Tech LABORATORIES, INC.

60 PARSIPPANY BLVD. BOONTON, N. J.

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ALL SCHEMATICS ARE LOCATED IMMEDIATELY FOLLOWING THE LAST PAGE OF TEXT.

N O T I C E

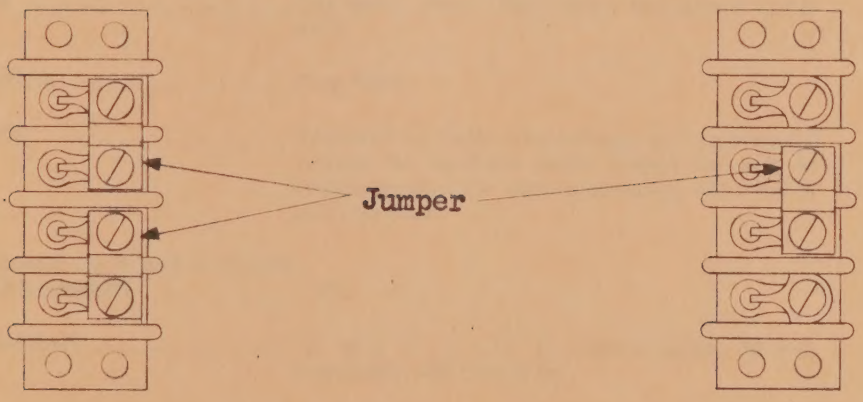
This instrument is designed to operate from a 117 volt A.C. or 234 volt A.C. power source. The nameplate located near the power cord indicates the voltage at which the instrument should be operated.

Should it be desired to change the input voltage!

1. Change transformer primary connections by moving the jumpers on the 4 terminal barrier strip located near the power cord (See figure below).

2. Change fuse (See figure below).

3. Reverse the nameplate located near the power cord.



For 117 Volt Operation

Fuse _____ Amp Slow Blow

For 234 Volt Operation

Fuse _____ Amp Slow Blow

MODEL 315
RESISTOR-NOISE TEST SET

GENERAL

The Model 315 is a compact, bench-size instrument for making precise quantitative measurements of excess resistor-noise. The index of measurement is microvolts-per-volt in a decade of frequency, expressed in decibels as recommended by the National Bureau of Standards.*

SPECIFICATIONS

RANGE ..... Resistor Test Range — 100 ohms to 22 megohms.
Noise Voltage — 0.6 microvolts in a decade to 1000 microvolts in a decade.
Applied DC Voltage — 3 to 300 volts.

FILTER ..... Geometric mean 1000 cycles, flat-topped bandpass.

DETECTOR ..... Pure RMS.

OUTPUT ..... Indicated for both noise voltage and applied DC voltage on separate panel meters as well as analog outputs for data processing. AC monitor jack included.

ACCURACY OF NOISE VOLTAGE MEASUREMENT ..... ±5%.

DIMENSIONS ..... 15" W x 11" H x 11" D overall, weight 30 lbs., completely self-contained.

POWER INPUT ..... 117 volts, 50-60 cycles.

*See: "A Recommended Standard Resistor Noise Test System," G. T. Conrad, Jr., N. Newman, & A. P. Stansbury; I.R.E. Transaction of the Professional Group on Components Parts; Vol. CP-7, No. 3, Sept. 1960. Reprint available from Quan-Tech Laboratories on request.

SECTION 1
OPERATING INSTRUCTIONS

1. For initial operation of the Resistor-Noise Test Set, plug the power cord into a suitable source of power 105 to 125 volts 50-60 cycles. Turn the switch to the ON position. Since there are vacuum tubes contained in this instrument, a few minutes warm-up will be required.

2. Insert the resistor to be tested in the test jig, making sure that firm contact is made with the resistor lead. Before closing the top, make sure that the voltage to be applied to the resistor is not excessive.

3. Set RESISTOR RANGE control to the value indicated for the test resistor.

4. Set the FUNCTION switch to CALIBRATE. Close the test jig top and adjust the CALIBRATE ADJUST control until the NOISE meter reads at the CAL line (10 db). Switch the FUNCTION switch to SYSTEM NOISE and note noise reading on NOISE meter. If necessary, adjust METER MULTIPLIER to give an on scale reading.

5. Switch FUNCTION switch to TOTAL NOISE and adjust VOLTAGE ADJUST to appropriate voltage. Typical voltage values can be obtained from Table 1-1.

6. To obtain the Noise Index, microvolts per volt in a decade, merely subtract the dc voltage reading (D) in db from the total noise voltage reading (T) in db. For resistor having small amounts of noise, that is, a small difference between system noise (S) and total noise, a correcton factor f(T—S), must be applied. This correction factor may be obtained from the graph in Figure 1-2. Thus the Index = T — f(T—S) — D in db.

7. Noise Index in db may be converted to noise voltage in microvolts/volt by referring to Figure 1-1.

MODEL 315 RESISTOR-NOISE TEST SET

Instructions for measuring noise in Back-Biased Diodes

1.) Turn on instrument and allow to warm up for a few minutes.

2.) Place short circuit in test jig, set FUNCTION switch to CALIBRATE, and adjust NOISE meter to red line. Remove short circuit.

3.) Set RESISTOR RANGE switch to 250K-22M position. Set FUNCTION switch to SYSTEM NOISE, close cover, and note NOISE reading (Should be somewhat less than 18db).

4.) Set FUNCTION switch to TOTAL NOISE and set VOLTAGE ADJUST to desired voltage.
NOTE: Minimum supply voltage in the Model 315 is 20-30 volts. For measuring diodes with a breakdown voltage below 30 volts, use the Model 314 Diode Noise Analyzer.

5.) Insert diode in test jig and close cover. Note noise reading. If there is less than 10db difference from the value noted in Step 3, use Correction Chart for presence of System Noise, as with resistors.

6.) If DC Voltage reading changes from Step 4, it indicates leakage in the diode, which may be calculated as the difference between the two voltage readings through the 1 megohm isolating resistor.

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MODEL 315 [illegible] TEST SET

Instructions for measuring noise in Zener Diodes

1.) Turn on instrument and allow to warm up for a few minutes.

2.) Place short circuit in test jig, set FUNCTION switch to CALIBRATE and adjust NOISE meter to red line.

3.) Set FUNCTION switch to SYSTEM NOISE and note NOISE meter reading. (Should be somewhat less than -8 db). Remove the short circuit.

4.) Select the appropriate position on the RESISTOR RANGE switch.

Resistor Range	Internal Isolating Resistor	Zener Current
0.1K- 2.5K	1K	10-30 ma
2.5K- 25K	10K	3-10 ma
25K- 250K	100K	0.3-3 ma
250K -22M	1Meg	30-300 μa

5.) Calculate the required DC supply voltage, which equals the Zener voltage plus the Zener current times the isolating resistor. VDC = Vz + Iz x Ris (For example, a 6 volt Zener with 5 ma through it. 6 + (5x10-3) (10 4) = 56 VDC). CAUTION! Do not exceed 1 watt dissipation in the isolating resistor.

6.) Close the cover and set the FUNCTION switch to TOTAL NOISE. Set the VOLTAGE ADJUST TO THE VALUE calculated in Step 4. NOTE: The minimum supply voltage in the Model 315 is 20-30 volts. It may be impossible with low voltage Zeners to obtain currents in the 10 to 30μa range. Use the Model 314 Diode Noise Analyzer for such measurements.

7.) Insert the Zener Diode and close the cover. The DC Voltmeter will read the Zener voltage, and the Noise meter the Zener noise. If there is less than 10db difference from the value noted in Step 3, use correction chart for presence of System Noise, as with resistors.

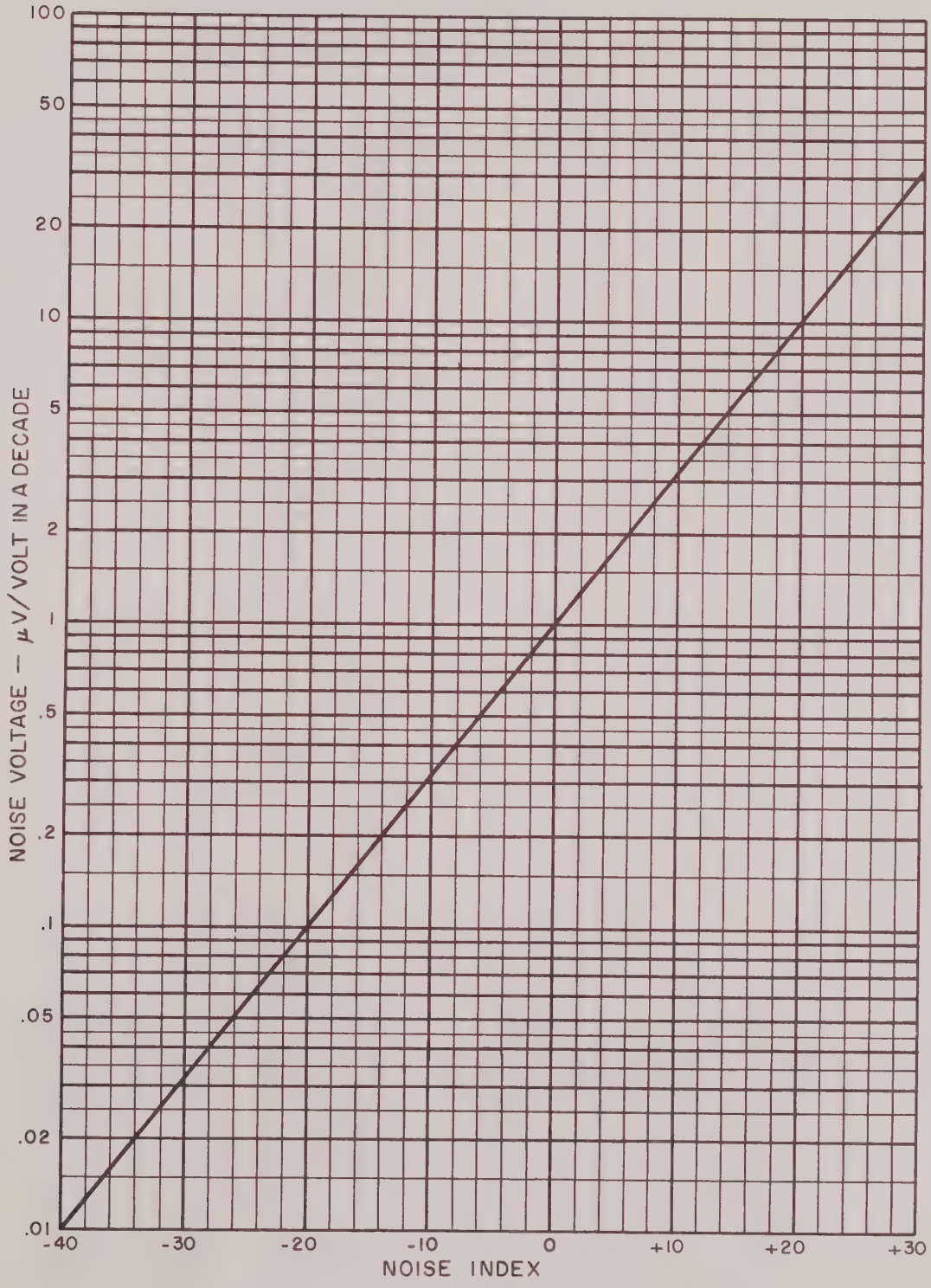


FIGURE 1-1. NOISE CONVERSION GRAPH

TABLE 1-1. RECOMMENDED OPERATING CONDITIONS

R Ω	R_m Ω	½ watt and High Wattage Ratings (D)* db	V Volts	P_{dc} MW	⅛, ¼ and 1/10 Watt Ratings (D)* db	V Volts	P_{dc} MW
100	1 K	10.1	3.2	100	10.1	3.2	100
120	1 K	11.6	3.8	120	10.9	3.5	100
150	1 K	13.5	4.7	150	11.8	3.9	100
180	1 K	15.1	5.7	180	12.5	4.2	100
220	1 K	16.9	7.0	220	13.4	4.7	100
270	1 K	18.3	8.2	250	14.3	5.2	100
330	1 K	19.2	9.1	250	15.1	5.7	100
390	1 K	19.9	9.9	250	15.8	6.2	100
470	1 K	20.7	10.8	250	16.7	6.9	100
560	1 K	21.4	11.8	250	17.5	7.5	100
680	1 K	22.3	13.0	250	18.3	8.2	100
820	1 K	23.1	14.3	250	19.2	9.1	100
1.0 K	1 K	24.0	15.8	250	20.0	10.0	100
1.2 K	1 K	24.8	17.3	250	20.8	11.0	100
1.5 K	1 K	25.8	19.4	250	21.7	12.2	100
1.8 K	1 K	26.6	21.2	250	22.5	13.4	100
2.2 K	1 K	27.4	23.4	250	23.4	14.8	100
2.7 K	10 K	28.3	26.0	250	24.3	16.4	100
3.3 K	10 K	29.2	28.7	250	25.2	18.2	100
3.9 K	10 K	29.9	31.2	250	25.9	19.7	100
4.7 K	10 K	30.8	34.3	250	26.7	21.7	100
5.6 K	10 K	31.5	37.4	250	27.5	23.7	100
6.8 K	10 K	32.3	41.2	250	28.3	26.1	100
8.2 K	10 K	33.2	45.3	250	29.1	28.6	100
10 K	10 K	34.0	50.0	250	30.1	32.0	100
12 K	10 K	34.8	54.8	250	30.9	35.0	100
15 K	10 K	35.8	61.2	250	31.8	39.0	100
18 K	10 K	36.6	67.1	250	32.5	42.0	100
22 K	10 K	37.4	74.2	250	33.4	47.0	100
27 K	100 K	38.3	82.2	250	34.3	52.0	100
33 K	100 K	39.2	90.8	250	35.1	57.0	100
39 K	100 K	40.0	98.7	250	35.8	62.0	100
47 K	100 K	40.7	108	250	36.7	69.0	100
56 K	100 K	41.5	118	250	37.5	75.0	100
68 K	100 K	42.3	130	250	38.3	82.0	100
82 K	100 K	43.1	143	250	39.2	91.0	100
100 K	100 K	44.0	158	250	40.0	100	100
120 K	100 K	44.8	173	250	40.8	110	100
150 K	100 K	45.8	194	250	41.7	122	100
180 K	100 K	46.5	212	250	42.5	134	100
220 K	100 K	47.5	234	250	43.4	148	100
270 K	1 M	38.6	85.0	26.8	38.6	85.0	26.8
330 K	1 M	40.0	99.0	29.7	40.0	99.0	29.7
390 K	1 M	41.0	112	32.2	41.0	112	32.2
470 K	1 M	42.1	127	34.3	42.1	127	34.3
560 K	1 M	43.1	143	36.5	43.1	143	36.5
680 K	1 M	44.2	161	38.1	44.2	161	38.1
820 K	1 M	45.1	180	39.5	45.1	180	39.5
1.0 M	1 M	46.0	200	40.0	46.0	200	40.0
1.2 M	1 M	46.8	218	39.6	46.8	218	39.6
1.5 M	1 M	47.6	240	38.4	47.6	240	38.4
1.8 M	1 M	48.0	250	34.7	48.0	250	34.7
2.2 M	1 M	48.0	250	28.4	48.0	250	28.4
2.7 M	1 M	48.0	250	23.2	48.0	250	23.2
3.3 M	1 M	48.0	250	18.9	48.0	250	18.9
3.9 M	1 M	48.0	250	16.0	48.0	250	16.0
4.7 M	1 M	48.0	250	13.3	48.0	250	13.3
5.6 M	1 M	48.0	250	11.2	48.0	250	11.2
6.8 M	1 M	48.0	250	9.2	48.0	250	9.2
8.2 M	1 M	48.0	250	7.6	48.0	250	7.6
10 M	1 M	48.0	250	6.2	48.0	250	6.2
12 M	1 M	48.0	250	5.2	48.0	250	5.2
15 M	1 M	48.0	250	4.2	48.0	250	4.2
18 M	1 M	48.0	250	3.5	48.0	250	3.5
22 M	1 M	48.0	250	2.8	48.0	250	2.8

* $D = 20 \log_{10} V$

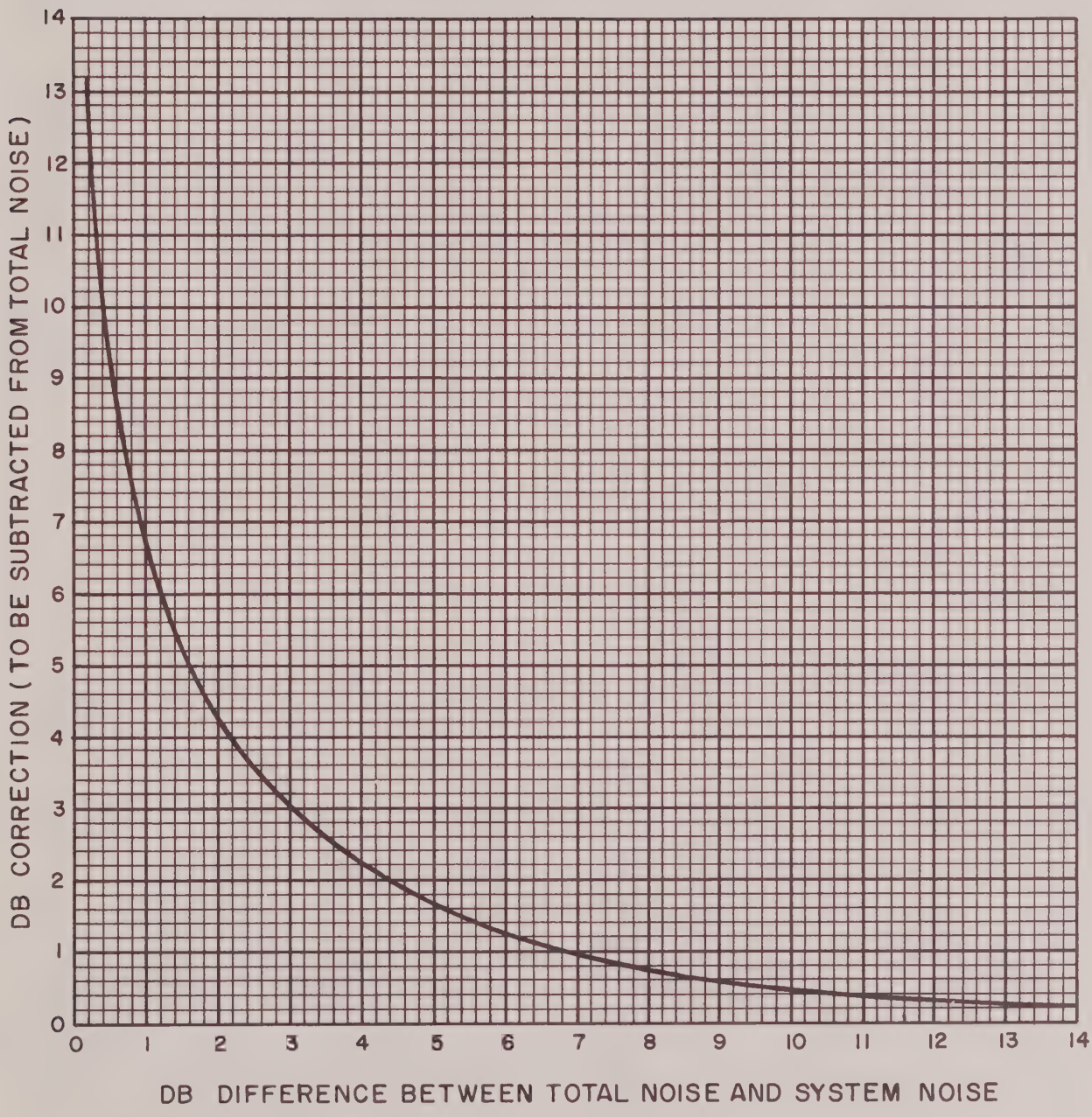


FIGURE 1-2. CORRECTION CHART FOR PRESENCE OF SYSTEM NOISE

SECTION 2

THEORY OF OPERATION

The Model 315 is completely self-contained and requires no accessories for operation. Basically, this instrument serves to measure the increase in noise due to current passing through a resistor.

The simplified block diagram of the system shown in Figure 2-1 illustrates the principles of operation. The dc supply supplies a variable dc voltage from 30 to ~~400~~ 500 volts to the test resistor and is isolated by the isolation resistors shown. The dc voltage across the resistor is measured with a vacuum tube voltmeter and the ac noise is amplified, filtered and metered with a thermal type meter. Since the test resistors are of different impedances, calibration of each different value may be made by adjustment of the system gain. Thus, the noise measured is equivalent to that which would come from a 1 ohm generator in series with a test resistor in the same manner as the calibration source.

Figure 2-2 of the detailed block diagram is an elaboration of the simplified diagram and is given in order to simplify servicing and location of parts in the equipment should it be required.

Figure 2-3, Typical Filter, is a plot of the four section Butterworth Bandpass Filter employed in the instrument. Slight deviations (in the order of a few percent) may be found from unit to unit. The filter has been designed with a center frequency of 1000 cycles and having a 1000 cycle bandpass.

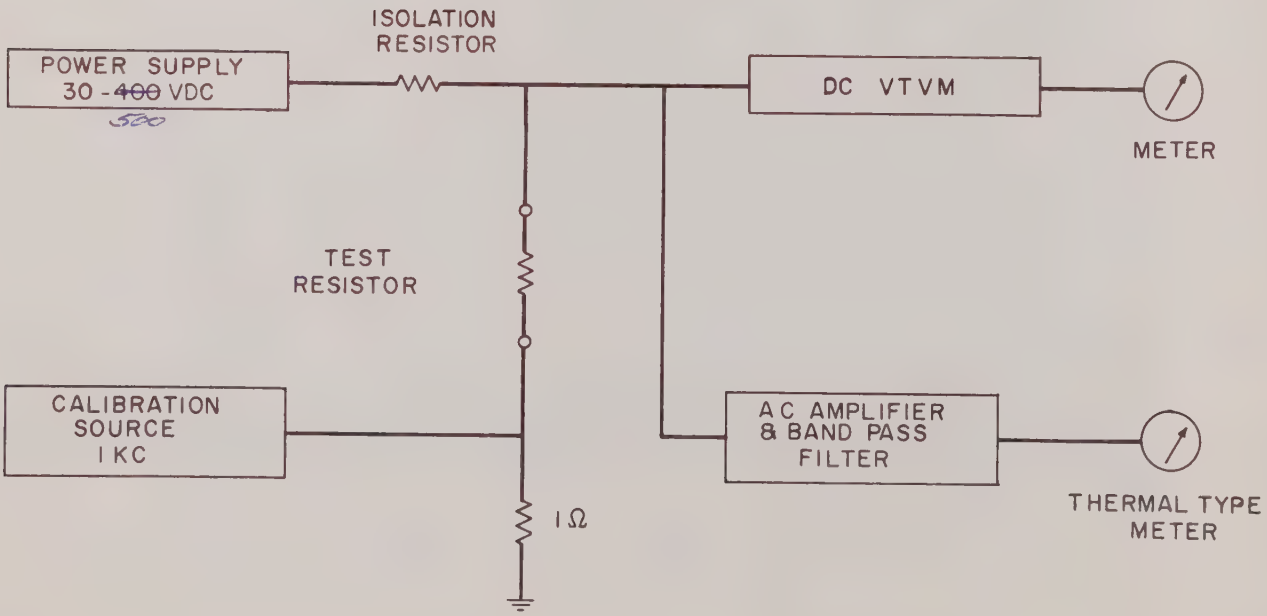


FIGURE 2-1. BLOCK DIAGRAM OF SYSTEM

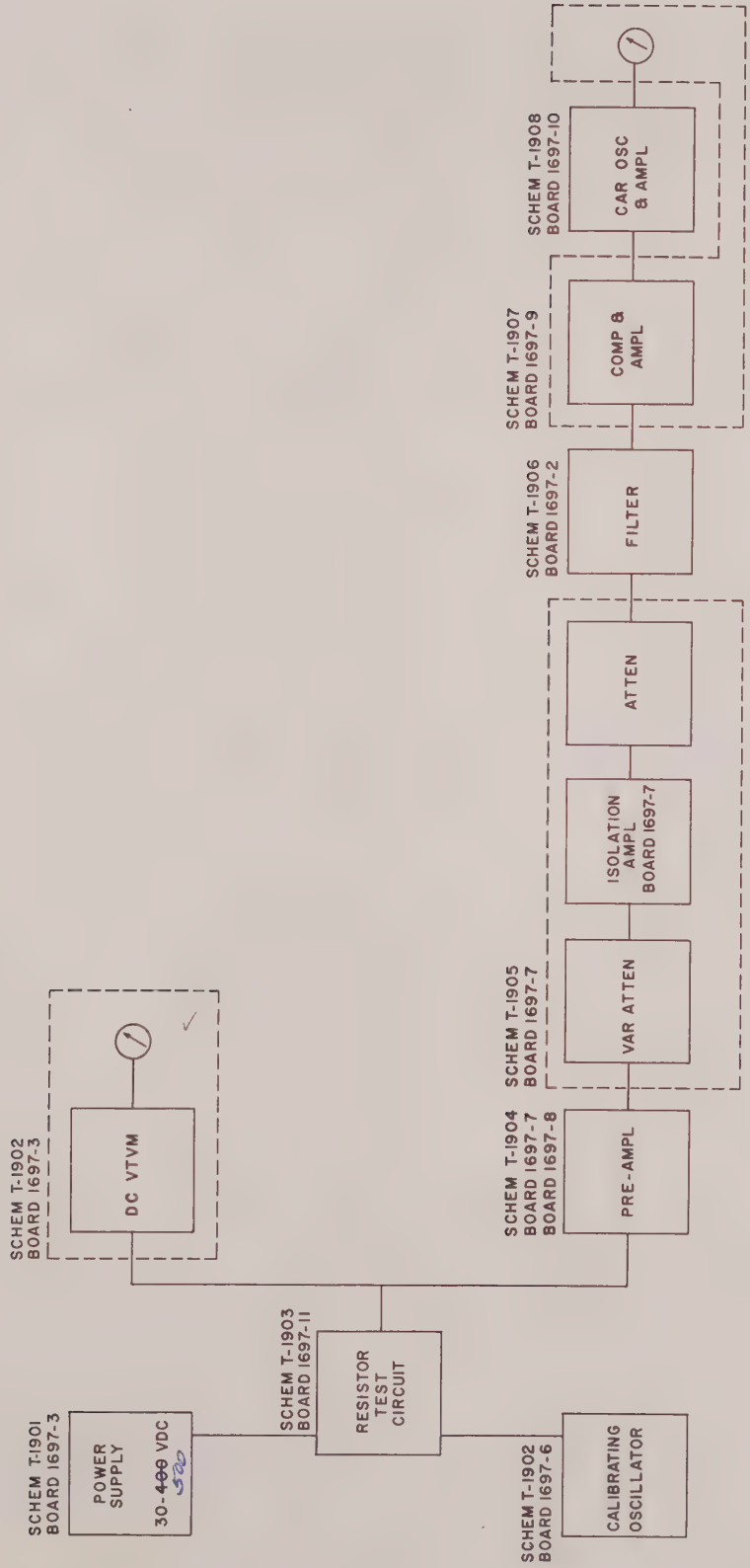


FIGURE 2-2. DETAILED BLOCK DIAGRAM — MODEL 315

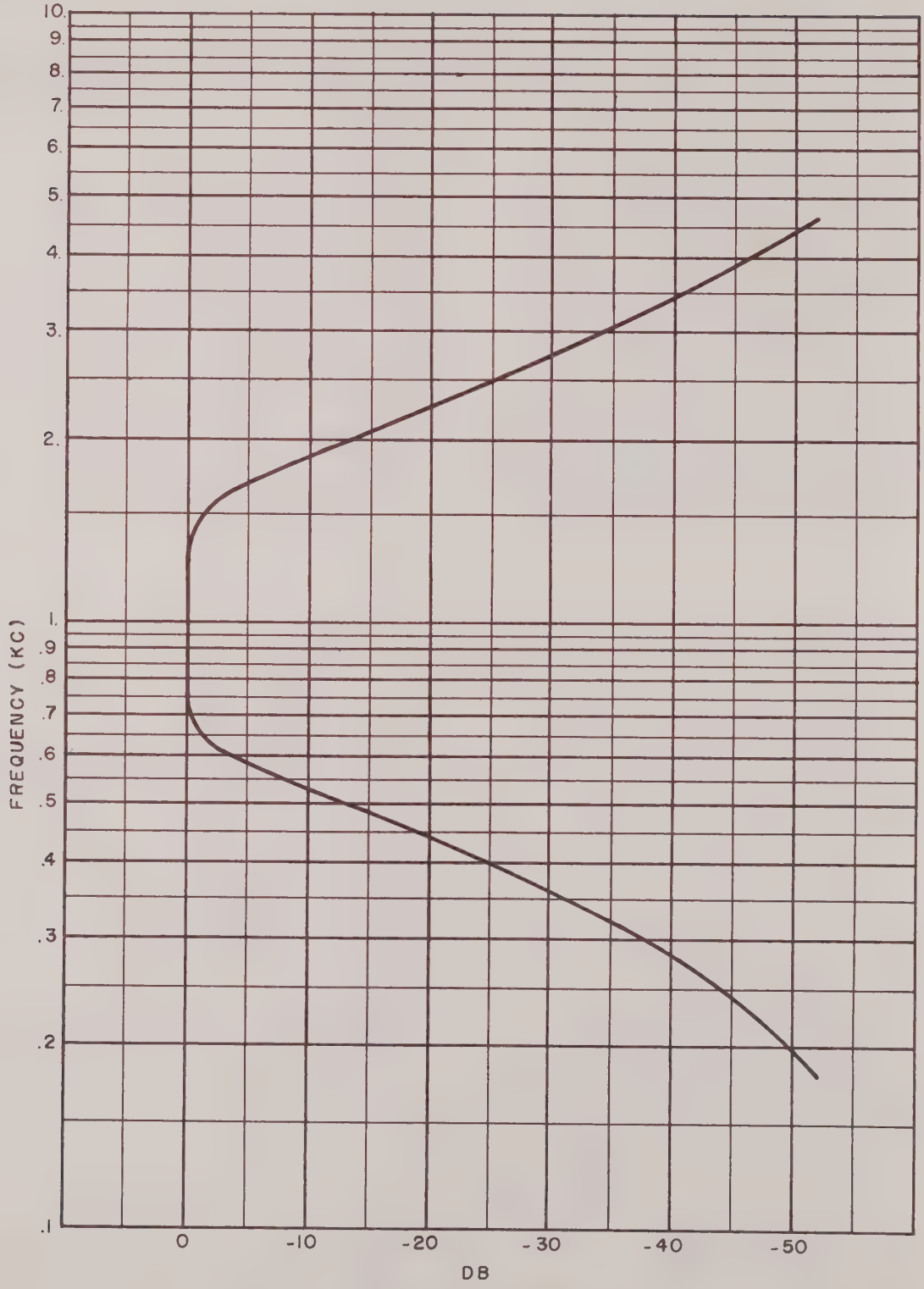


FIGURE 2-3. TYPICAL FILTER

SECTION 3

GENERAL CURRENT NOISE DISCUSSION

The limitation in sensitivity of electronic systems due to noise has been known for over 30 years. Even ideal amplifiers have a threshold determined by the resistance of the signal source. Nyquist's formula, based on the statistical theory of thermodynamics, gives the noise voltage at the terminals of an impedance.

$$\overline{E_{th}^2}^{1/2} = (4k\,T\,B\,R)^{1/2} \qquad \text{eq. 1}$$

where $\overline{E_{th}^2}^{1/2}$ = rms value of the thermal noise voltage, in volts.
k = Boltzman's constant, 1.38×10^{-23} Joules/degree.
T = Temperature in degrees Kelvin.
B = Bandwidth of measuring system in cycles.
R = Real part of the impedance Z, in ohms.

Although most engineers are experienced with the noise generated in vacuum tubes, few have an understanding of the types or magnitudes of noise generated in other electronic devices such as, transistors and other semi-conductors, electrolytic capacitors, and current carrying resistors. Perhaps one of the least realized, but important source of noise, is that of current carrying resistors.

If current is passed through a resistor, an increase in noise above the thermal noise occurs. In wire wound, and some metal film resistors, this effect is of second order and may not be detectable unless the resistors are faulty. However, in resistors, such as composition, deposited carbon resistors, and certain metal film resistors, the effect is by no means negligible and may exceed the theoretical value by an order of magnitude or more. The use of such noisy resistors in critical points of low level circuits may increase the noise at the input by the same factor.

Unlike thermal noise, which is "flat" or independent of frequency (well into the gigacycle region), resistor current noise has a decreasing power vs. frequency spectrum. The relationship of frequency, dc current, and noise in a resistor is given in the equation,

$$\overline{e\,(f)^2} = C\,I^{\alpha} f^{-\gamma} \qquad \text{eq. 2}$$

where $\overline{e\,(f)^2}$ = Mean square noise voltage per cycle of bandwidth at a frequency f.
C = Constant dependent upon the noise quality of the resistor.
I = dc current in amps.
f = Frequency in cycles.
α = 1.4 to 2.2 dependent upon the resistor, and has a mean value of 1.9.
$\gamma \cong 1.0$.

Thus, for most purposes equation 2 may be written,

$$\overline{e\,(f)^2} = C\,I^2 f^{-1} \qquad \text{eq. 3}$$

showing the $1/f$ characteristics of noise power spectral density.

The resistor current noise for a given bandwidth may be obtained by integration of $e\,(f)^2$

$$\overline{E_c^2} = \int \overline{e\,(f)^2}\,df \qquad \text{eq. 4a}$$

Since $\overline{e\,(f)^2} = CI^2 f^{-1}$, then

$$\overline{E_c^2} = \int CI^2 f^{-1}\,df \qquad \text{eq. 4b}$$

$$\overline{E_c^2} = CI^2 \ln f \Big|_{f_1}^{f_2} \qquad \text{eq. 4c}$$

$$\overline{E_c^2}^{1/2} = \left[CI^2 (\ln f_2 - \ln f_1) \right]^{1/2} \qquad \text{eq. 4d}$$

or

$$\overline{E_c^2}^{1/2} = \left[CI^2 \ln\left(\frac{f_2}{f_1}\right) \right]^{1/2} \qquad \text{eq. 4e}$$

It can be seen from Equation 4e that bandwidths having equal ratios of upper to lower cut-off frequencies, such as octaves or decades of bandwidth, the current noise magnitudes are equal.

SECTION 3 - GENERAL CURRENT NOISE DISCUSSION

The National Bureau of Standards has recommended the unit microvolts per volt in a decade in db as a standard for current noise in a resistor. There are two advantages of this unit; one, the ease of measurement; and two, the familiarity of the terms for the engineer.

The above unit or INDEX is defined as:

$$\text{INDEX} = 20 \log \frac{\overline{E_c^2}^{1/2}}{V} \text{ db (in a decade of frequency)} \qquad \text{eq. 5}$$

Therefore equation 4e may be changed to

$$\overline{E_c^2}^{1/2} = V\, 10^{\text{INDEX}/20} \left(\log \frac{f_2}{f_1} \right)^{1/2} \qquad \text{eq. 6}$$

where $\overline{E_c^2}^{1/2}$ = total rms current voltage.
V = dc voltage.
f_1 & f_2 = lower and upper effective bandpass limits.

Since current noise is of a $1/f$ nature, its effect will be masked by thermal noise at higher frequencies. In practice, this is usually true above 100 kc. Both thermal noise and current noise have a Gaussian or normal voltage amplitude distribution. Since their sources are independent they may be added as shown,

$$\overline{E_{Total}^2}^{1/2} = \left[\overline{E_c^2} + \overline{E_{th}^2} \right]^{1/2} \qquad \text{eq. 7}$$

Figure 3-1 shows the noise spectral density as a function of frequency of a typical 10 K ½ watt resistor with 20 volts dc applied.

This resistor has a current noise INDEX of approximately —10 db (0.3 μv/v in a decade). It has been found from the evaluation of a large quantity of composition resistors that nearly all fall in the range of 0.1 to 10 μv/v in a decade of current noise.

We may add the noise in decades in an r m s fashion. That is, for 2 decades of $1/f$ noise we have $\sqrt{2}$ times as much noise voltage, and for 3 decades $\sqrt{3}$ times as much. It should be pointed out that we have not added equal bandwidths but equal percentage bandwidths. Thus, for bandwidths other than one decade equation 6 must be used.

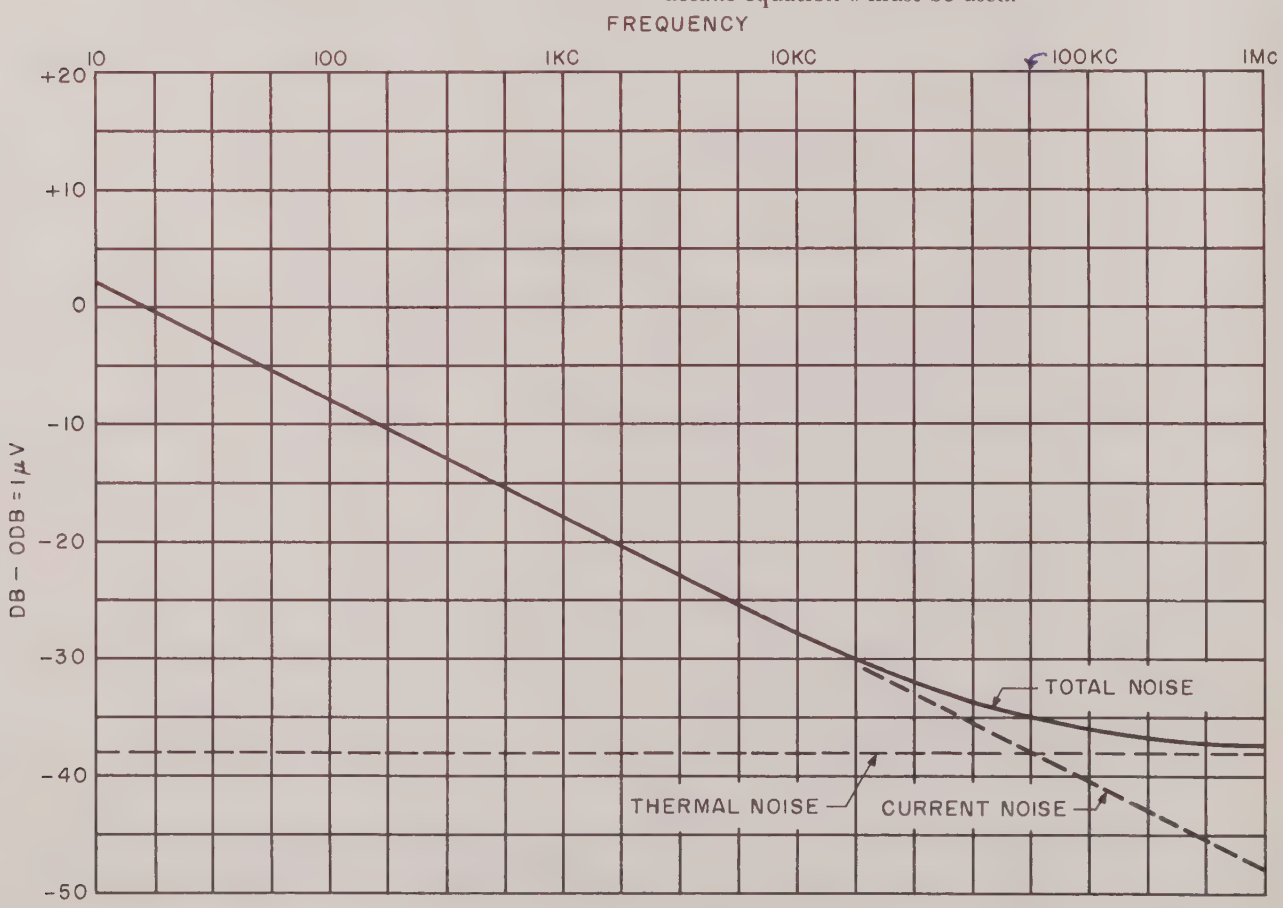


FIGURE 3-1. NOISE VOLTAGE SPECTRAL DENSITY 10K OHM COMPOSITION RESISTOR, 20 VDC

APPLICATIONS:

In order to apply the Index of current noise of a resistor in circuit design one must know the Index, the voltage and the position in the circuit. Let us look at a few practical applications of this noise. Since composition resistors of reasonable quality have an Index reasonably close to 0 db in a decade, we can see that the current noise voltage is more a function of the applied dc voltage than of the value of the resistor.

Assume a 10 K resistor has an Index of 0 db. This resistor will appear as a noise generator of 10 K impedance and for every volt of dc applied to its terminals the generator will develop 1 μv of current noise in one decade of bandwidth, or, in a bandwidth from 10 cycles to 10 kc (three decades) 1.73 μv/v of noise. For ten volts dc it would generate 17.3 μv of noise. By comparison, the thermal noise over the same bandwidth would be 1.25 μv. If this resistor were the bias resistor of a transistor input stage as shown in Fig. 3-2(a) the noise would be 14 times the theoretical minimum.

The effects of current noise are quite noticeable even in the plate circuits of vacuum tube amplifiers, and to a lesser extent in transistor collector circuits. The 100K resistor in Fig. 3-2(b) having a noise Index of 0 db and 100 volts applied voltage would result in approximately 5.5 μv of noise equated to the input, or an equivalent input resistance of 0.16 megohms. In this application negative feedback would reduce the noise in proportion to the feedback.

Similar results will be found in operational amplifier feedback, resistors, power supply references, servo loops, etc.

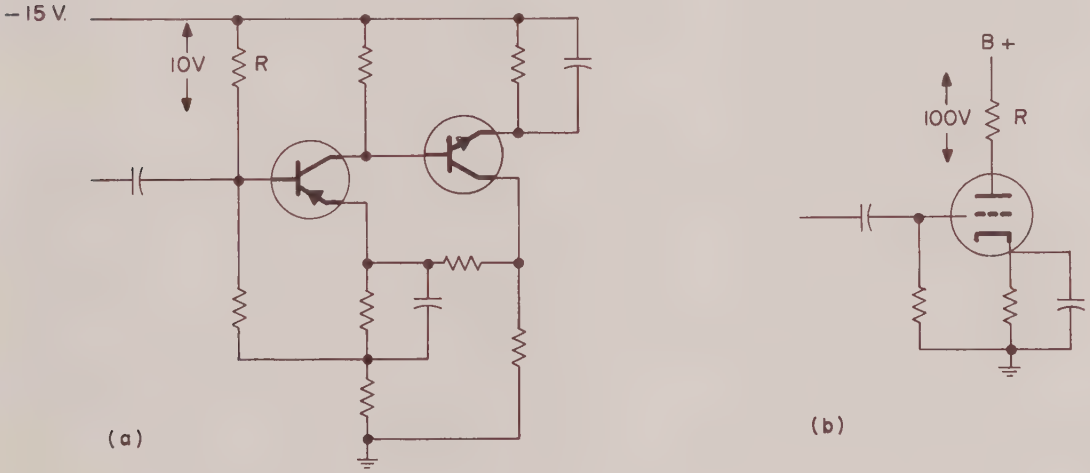


FIGURE 3-2. TYPICAL RESISTOR CURRENT NOISE SOURCES

SECTION 4
CIRCUIT ANALYSIS

POWER SUPPLY
(Board No. 1697-3) (Schematic T-1901)

The power supply shown on Schematic T-1901 supplies all regulated dc for the instrument. The reference for these supplies is the Zener Diode CR112. Transistors Q104, Q105, Q106 and Q107 form a double differential amplifier. Q106 drives the base of Q103 which is the pass transistor and supplies —26 volts for the transistor amplifiers, filters, measuring circuit and the filaments of all vacuum tubes except V101. Q101 and Q102 function as a sub-supply for +26 volts used in the vacuum tube voltmeter. The reference for this sub-supply is formed by the voltage divider network R103 and R117. The variable output voltage (+30 to ~~+400~~ +358 volts) is obtained from V101 and is varied by R107 which in conjunction with R106 and R127 form a voltage divider and obtains its reference from ~~the —26v supply~~ CR112. Q109 and Q108 form an emitter coupled amplifier to drive V104, and in turn V101. The +250 volt and +150 volt output obtained from V102 and V103 supply voltage for the vacuum tube voltmeter and preamplifier tubes respectively.

VACUUM TUBE VOLTMETER
(Board No. 1697-3) (Schematic T-1902)

The vacuum tube voltmeter design consists of a balanced pair of tubes V202 and V203. The meter multiplier attenuator merely acts as a shunt at the grid and is in series with the 8 megohm resistor found in the resistor test circuit schematic. V201 is a protective diode to clamp severe overloads and keep them from reaching the transistorized portion of the circuit. Q201 and Q203 act as grounded collector impedance converters. Q202 operates as a differential detector and drives the output meter and dc analog output. The meter is adjusted to read full scale for 10 volts at the analog output.

CALIBRATING OSCILLATOR
(Board No. 1697-6) (Schematic T-1902)

The calibrating oscillator consists of a single transistorized transformer coupled oscillator with a resonant tank in the collector circuit. The magnitude of oscillation is controlled from the base tie point and may be adjusted over narrow limits by R221. The magnitude of the 1000 cycle signal at the output to the test circuit should be 650 microvolts. The calibrating oscillator circuit is separately regulated by CR204.

RESISTOR TEST CIRCUIT
(Schematic T-1903)

The sensitive portion of the resistor test circuit is enclosed in an iron shield below the chassis, and the position of leads in this area is critical.

PRE-AMPLIFIER
(Board No. 1697-7) (Schematic T-1904)

The entire pre-amplifier is doubly shielded, and well filtered supplies are required. The amplifier consists of V401 and V402 operating as cathode followers in parallel, thus reducing the equivalent noise resistance at the input. Adjustment of R401 will vary the plate voltage on the tubes to an optimum point, normally about 50 volts. Excessive voltage increases the noise, and insufficient voltage may cause grid current to flow and again result in excessive noise for high input impedance. Q401, Q402 and Q403 form a PNP-NPN-PNP amplifier having a large amount of feedback, and a gain of approximately 75.

ISOLATION AMPLIFIER AND ATTENUATOR
(Board No. 1697-7) (Schematic T-1905)

The purpose of the isolation amplifier is to provide isolation and gain between the variable gain adjust (CALIBRATE ADJUST) and the meter multiplier attenuator. It consists of a PNP-NPN-PNP feedback triple and has a gain of approximately 150.

FILTER AMPLIFIER
(Board No. 1697-2) (Schematic T-1906)

The filter consists of a 4 section Butterworth filter, and each filter section operates essentially in the same manner. The output impedance of each transistorized amplifier in the filter is quite low, and the input impedance is determined by the base biasing resistors of each section. The Q of each of the filter resonant sections is determined by the respective base biasing resistors. Thus, each section of the filter is isolated from the other, and gain is obtained without overload due to interference.

MEASURING CIRCUIT

Compressor and Amplifier (Board No. 1697-9) (Schematic T-1907)

Carrier Oscillator and Amplifier (Board No. 1697-10) (Schematic T-1908)

The entire meter amplifier consists of an input voltage amplifier Q701 and Q702, volume compressor CR701, CR702, CR703, CR704 and Q703, a power amplifier Q704, Q705, Q706 and Q707, a hot wire element R729 and a level detector. The principle of operation of this part of the circuitry is to provide the hot wire element R729 with a constant average amplitude signal. This signal results in heating of the resistive element. The resistance due to heating of the element is then detected by measuring the magnitude of a 100 KC carrier signal which is supplied to the hot wire element by Q803. Q802, Q801 amplify the carrier signal across the hot wire element. This signal is rectified by CR801 and CR803. A threshold detector and amplifier consisting of Q709 and Q708 amplify the error signal between the reference threshold (adjustable by R735) and in turn control the volume compressor to maintain a constant amplitude signal again at R729, the hot wire element. This signal should be in the order of 2 volts.

SECTION 5
SERVICING

A thorough understanding of the circuitry is required for proper servicing. However, the individual sections of the instrument are straightforward and are simple to service.

The tubes in the vacuum tube voltmeter are selected, and their replacement will require matching.

CAUTION: The power supply contains high voltages.

If components must be removed from the printed circuit boards, use an iron only hot enough to give a good rapid solder connection. *DO NOT* use a solder gun. NEVER apply pressure which will push the copper from the board.

SCHEMATIC LEGEND

K = 1000

⊘ = Screw Driver Adjustment

▭ = Front Panel Control

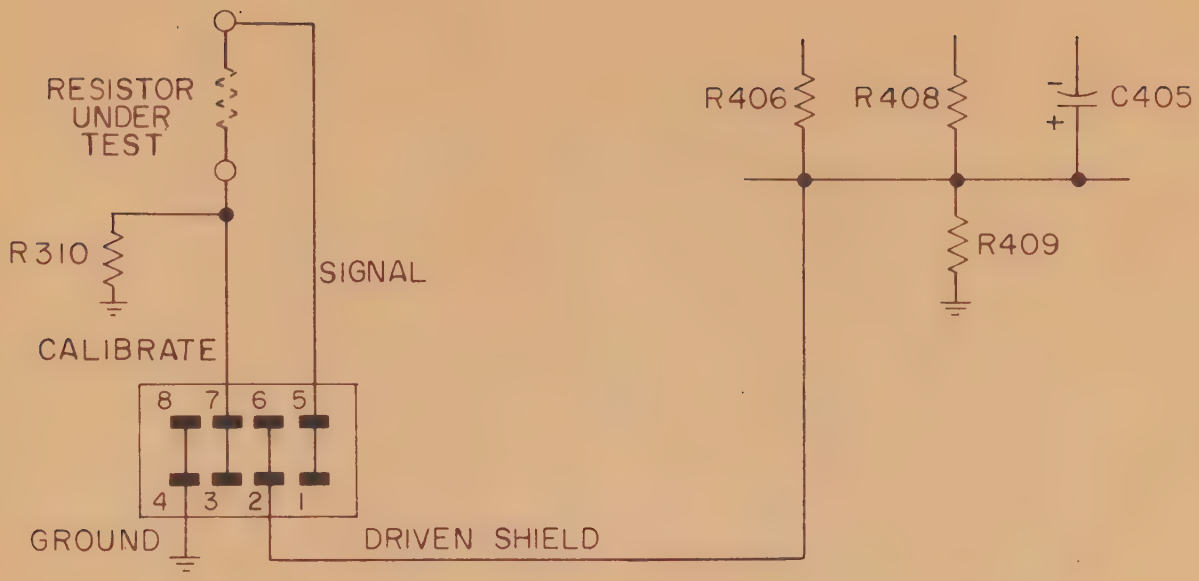
* = Factory Select

Unless Otherwise Specified:

All Resistances in Ohms

All Resistors ½ Watt

All Capacitances in Microfarads



SCHEMATIC – REMOTE OPERATING CABLE CONNECTOR

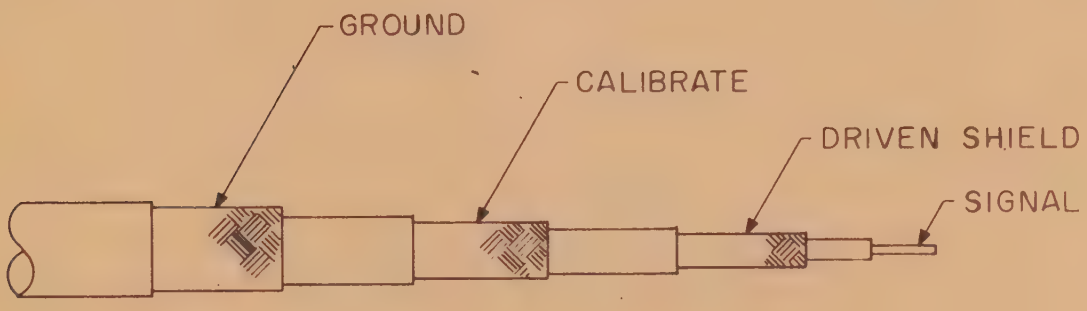
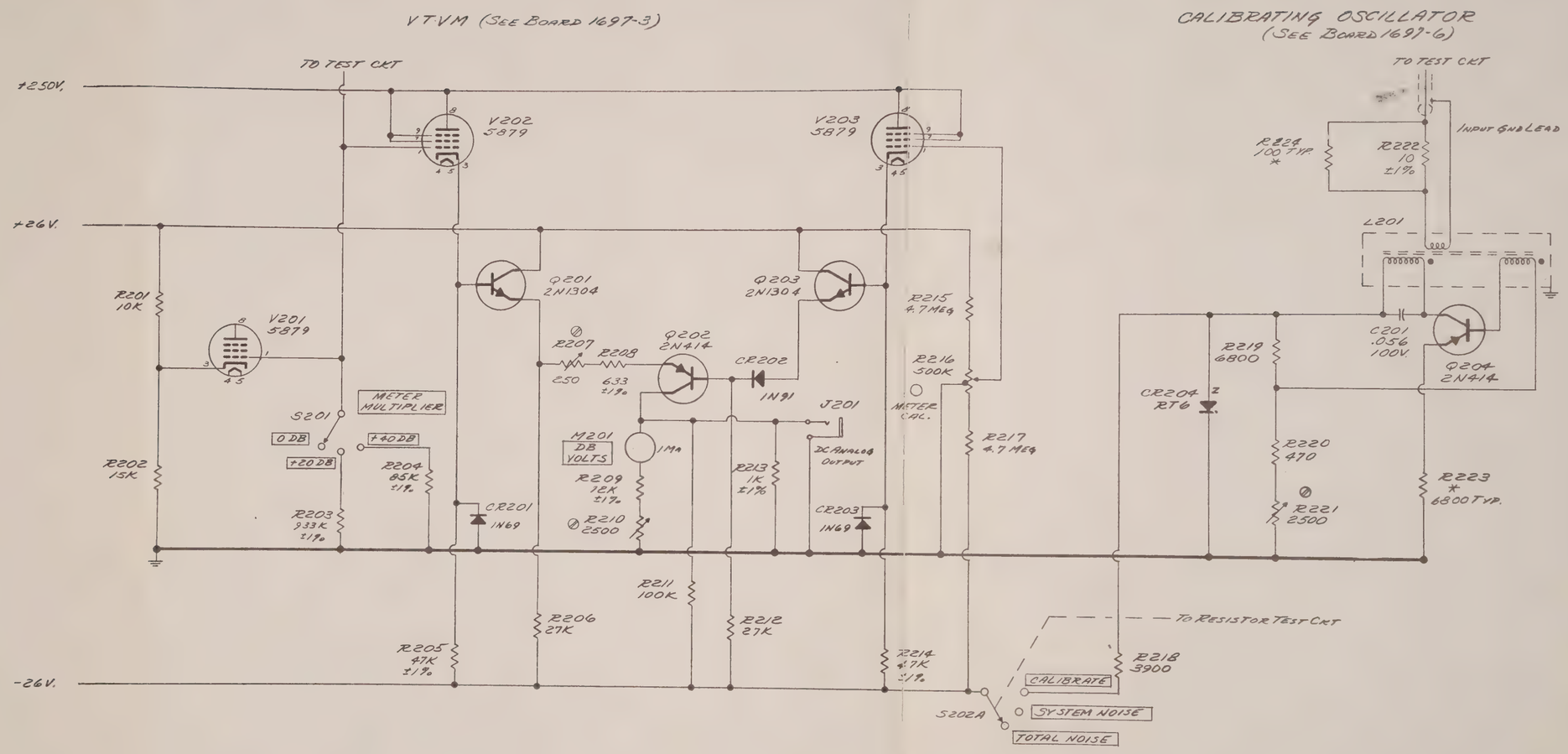
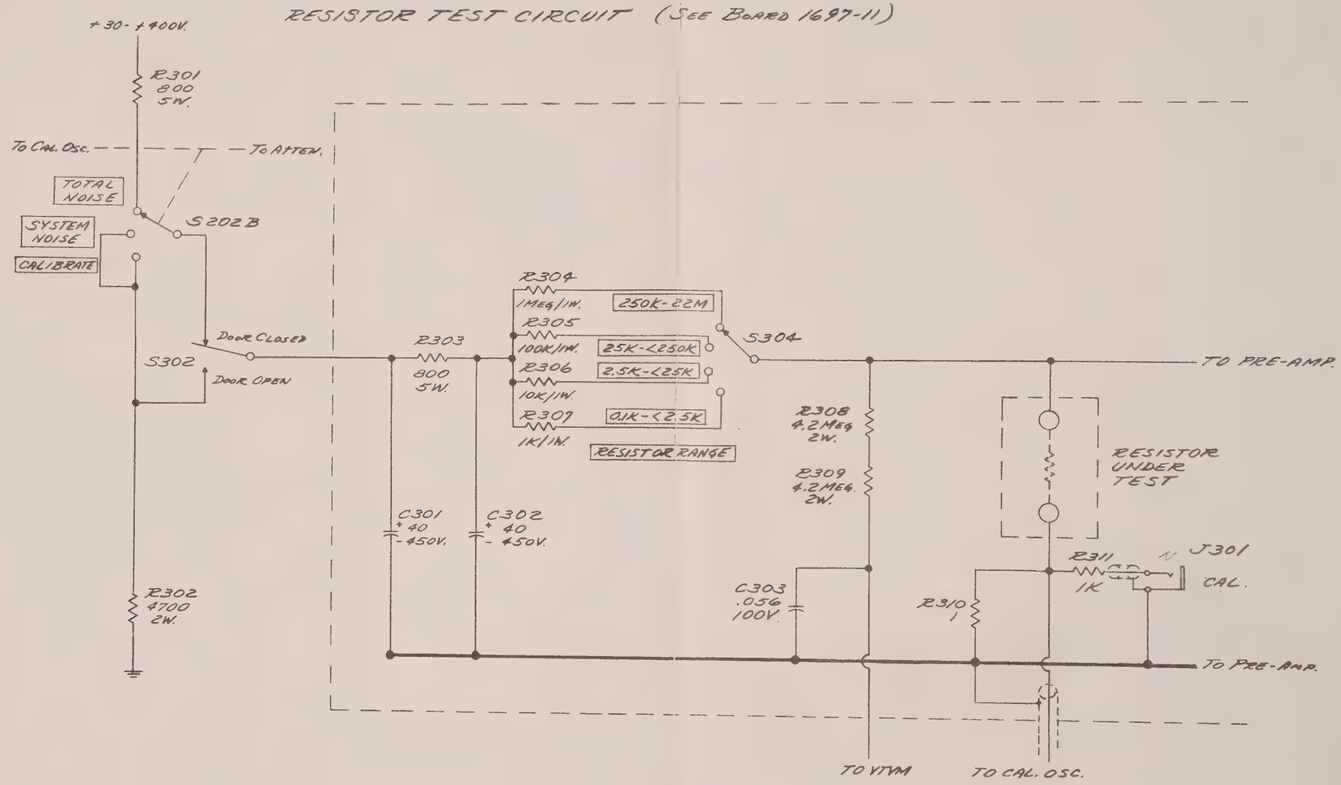


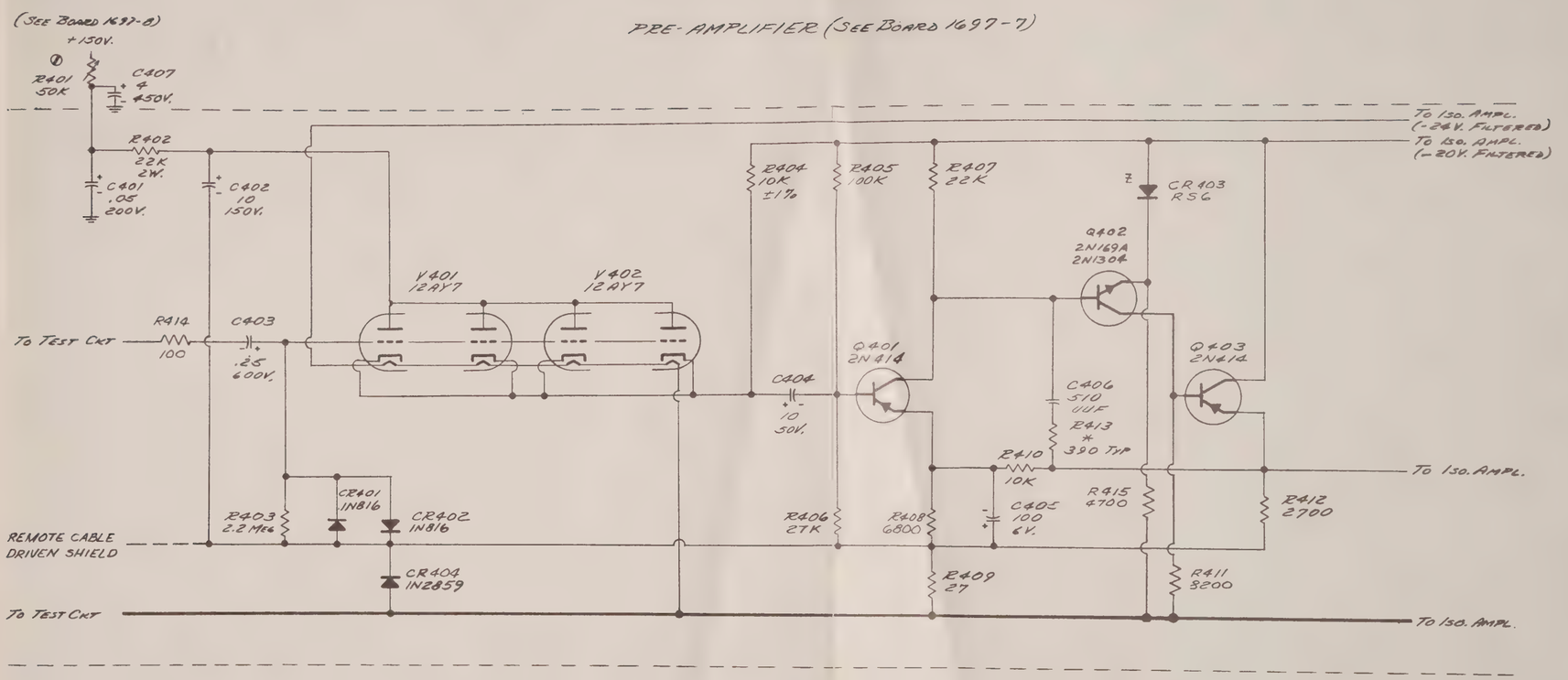
DIAGRAM – REMOTE OPERATING CABLE, TYPE 1107



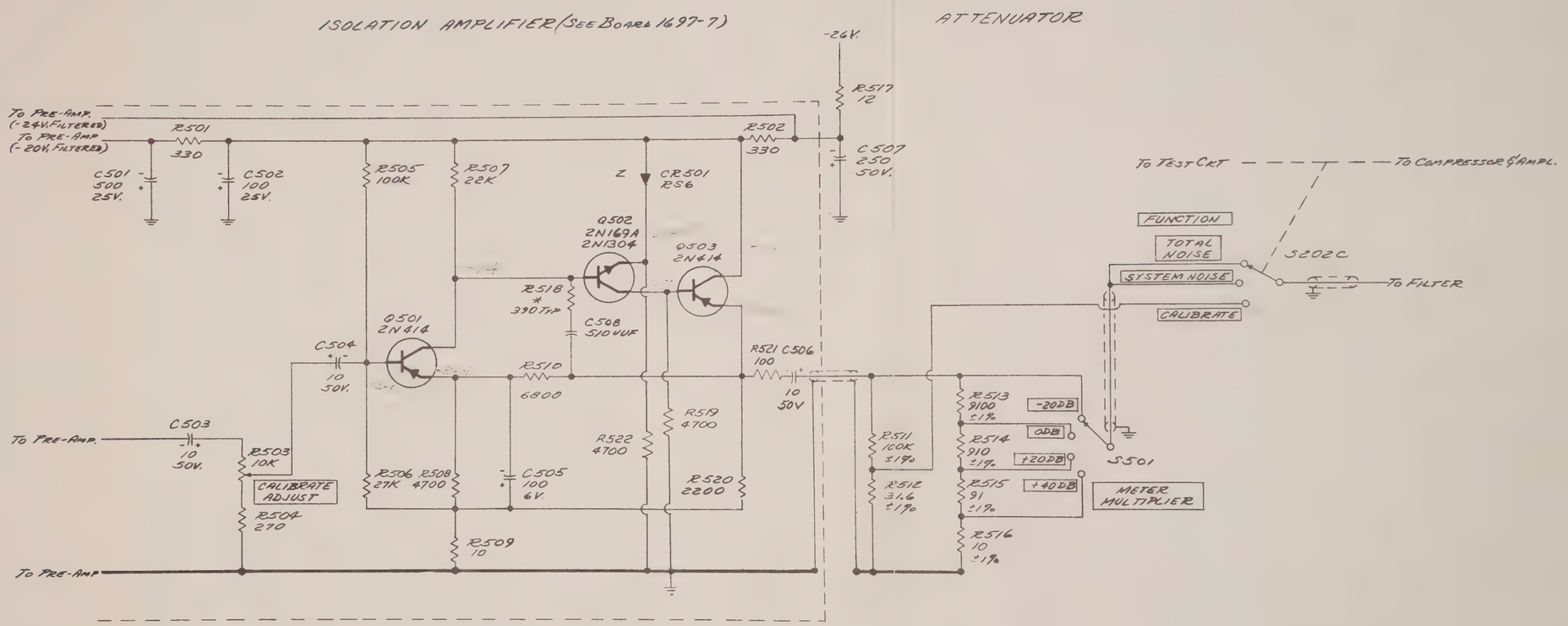
SCHEMATIC
VTVM AND CALIBRATING OSCILLATOR
MODEL 315
(Drawing No. T-1902-C)



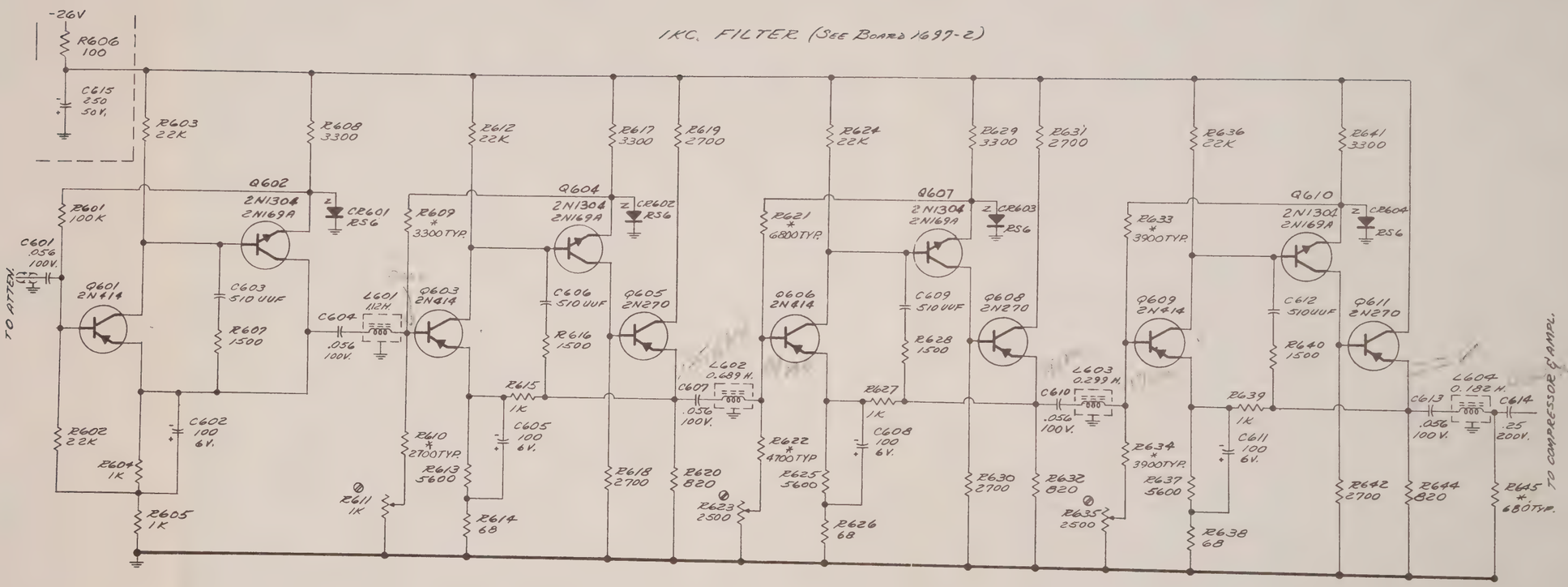
SCHEMATIC
RESISTOR TEST CIRCUIT
MODEL 315
(Drawing No. T-1903-C)



SCHEMATIC
PRE-AMPLIFIER
MODEL 315
(Drawing No. T-1904-G)

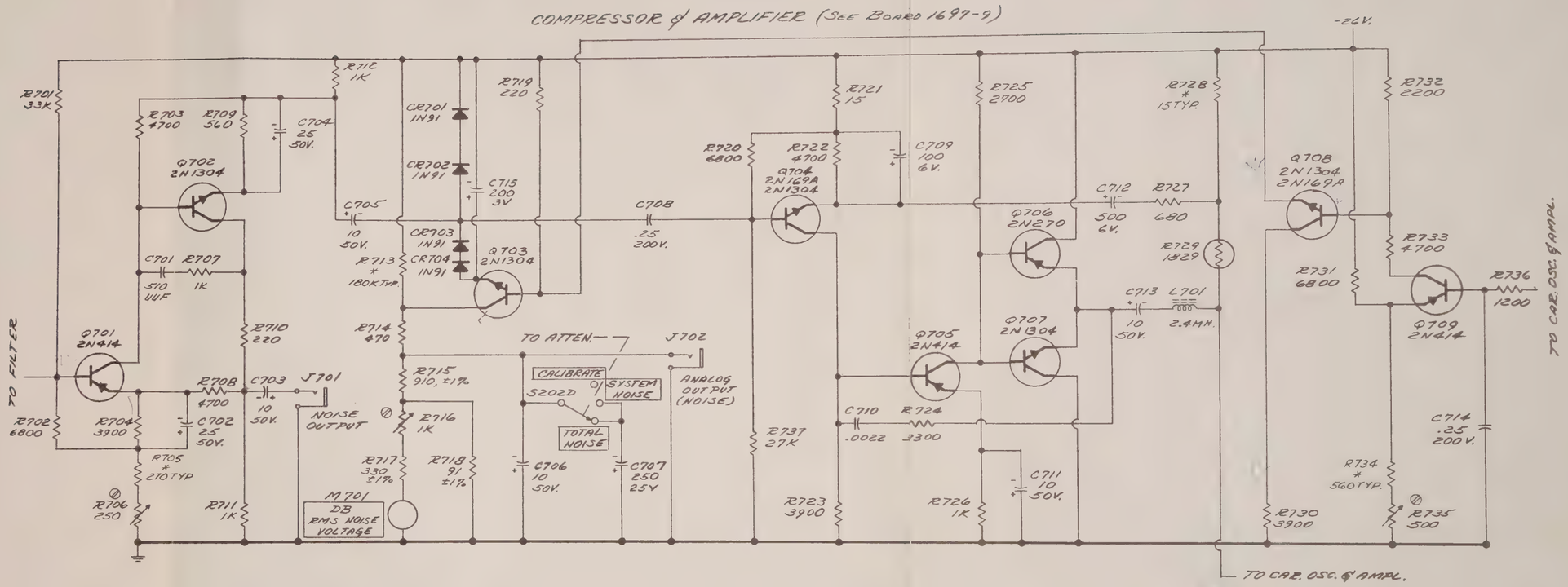


SCHEMATIC
ISOLATION AMPLIFIER AND ATTENUATOR
MODEL 315
(Drawing No. T-1905-F)

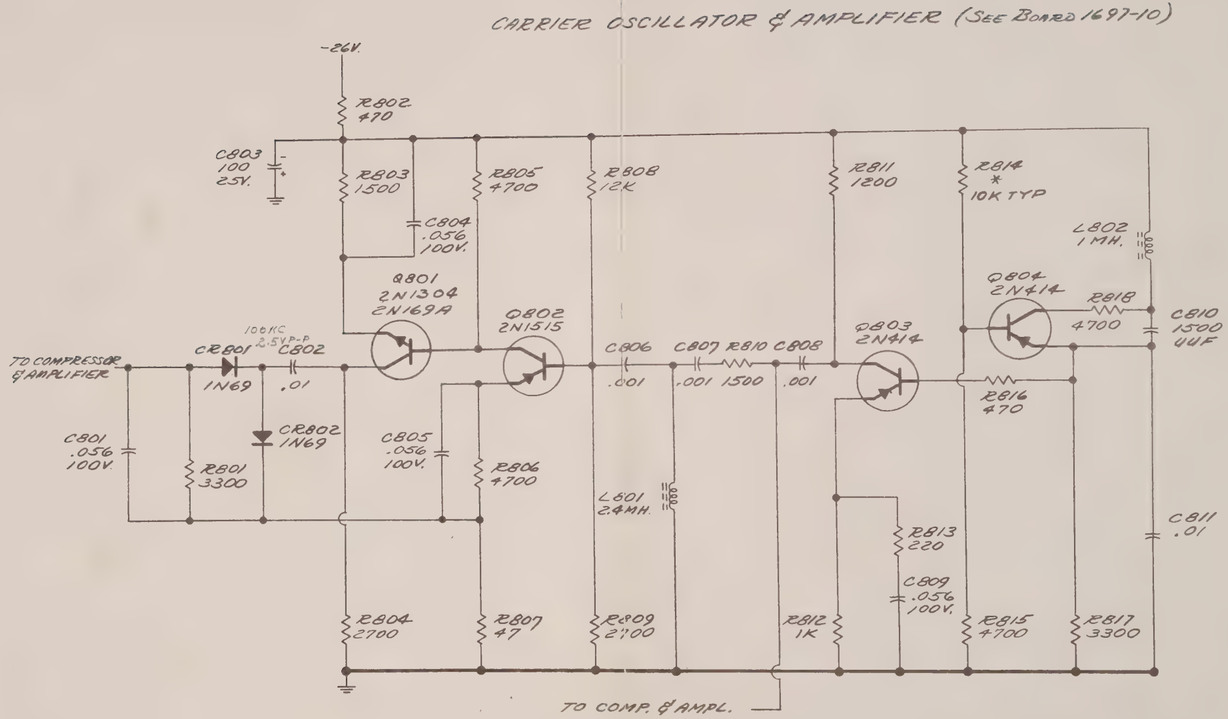


1KC. FILTER (See Board 1697-2)

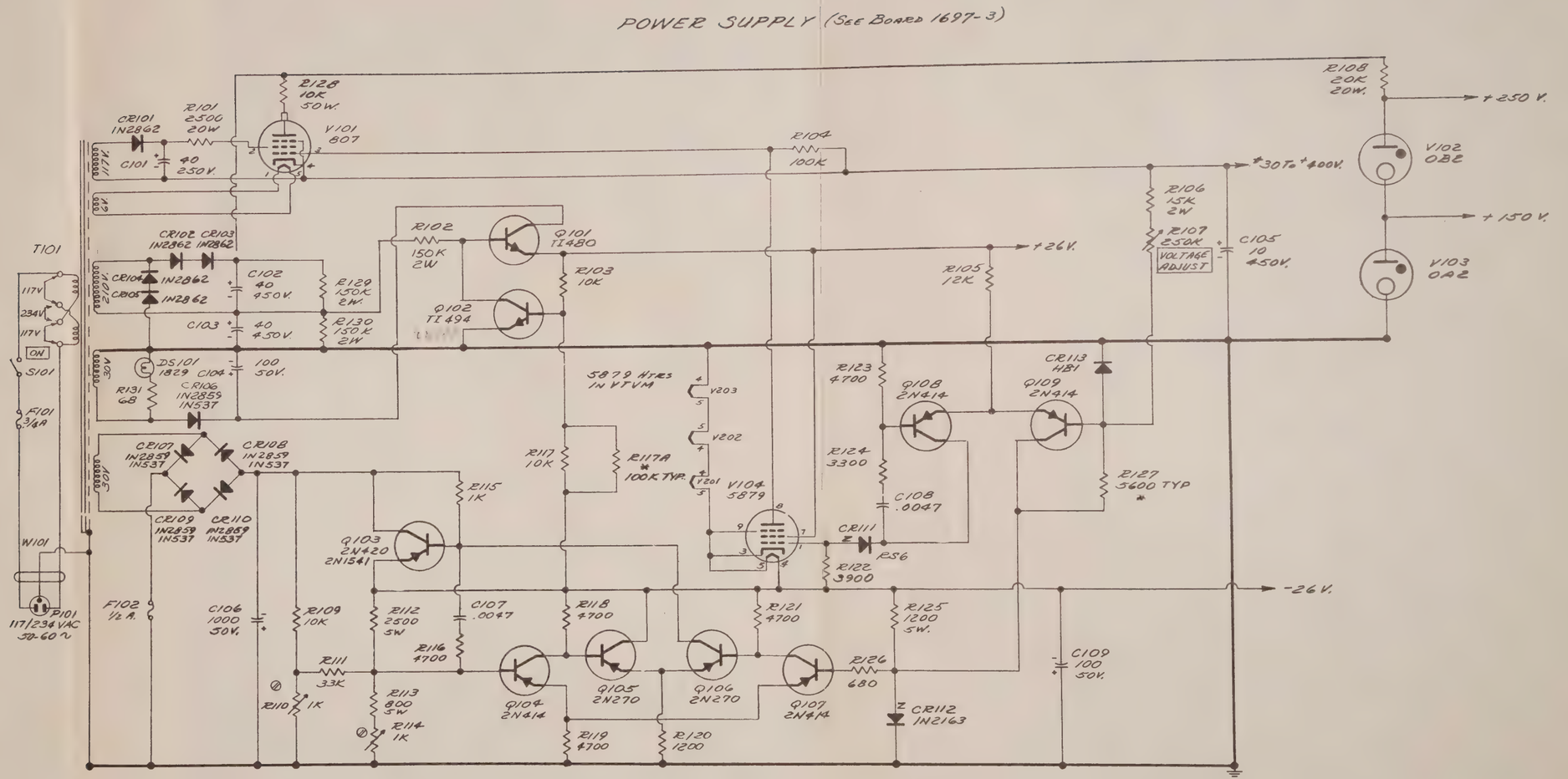
SCHEMATIC
FILTER
MODEL 315
(Drawing No. T-1906-F)



SCHEMATIC
COMPRESSOR AND AMPLIFIER
MODEL 315
(Drawing No. T-1907-F)



SCHEMATIC
CARRIER OSCILLATOR AND AMPLIFIER
MODEL 315
(Drawing No. T-1908-C)



SCHEMATIC
POWER SUPPLY
MODEL 315
(Drawing No. T-1901-G)

A Recommended Standard Resistor-Noise Test System*

G. T. CONRAD, JR.†, MEMBER, IRE, N. NEWMAN†, ASSOCIATE, IRE, AND A. P. STANSBURY‡

Summary—This paper describes a recommended standard resistor-noise test system, with the theory of operation given in considerable detail. The system provides for the measurement of a definitive resistor-noise quality *Index*. This *Index*, the "microvolts-per-volt" *Index*, is a modified form of the "conversion gain" index formerly used at NBS. The *Index* is a decibel expression for the number of microvolts of noise per volt of applied dc voltage transmitted in a frequency decade. A practical example is given to illustrate how the *Index* may be used to calculate the magnitude of current noise generated by fixed resistors in a given application. To complete the example, the combined effects of thermal noise, tube noise, and current noise are then considered.

A description is also given of an existing test set which is based upon specifications associated with the recommended measuring system. The full specifications for this set are given and distinctive design features are discussed. The simple operation of the test set and the computation of the *Index* are carefully explained.

I. Introduction

ALL resistors generate a fluctuation voltage, commonly called "thermal noise," whose rms value is predictable. Resistive elements having granular structures exhibit additional noise when carrying current. Defective wire-wound resistors also exhibit a similar increase in fluctuation voltage. This increase in fluctuation voltage is caused by current noise.[1] Unlike thermal noise, which is a function of resistance and temperature as given in Nyquist's equation,[2] no completely definitive relationship is known for current noise. Current noise has been observed to vary widely with such factors as geometry,[3] materials, and methods of construction. These factors vary for resistors of different power ratings and resistance values, and with different manufacturers. Therefore, the current-noise characteristics of groups of resistors must be established by measurement.

At present, there is no generally accepted standard method for measuring and evaluating resistor current noise. The noise measuring system described in this paper is intended to meet this need and is currently being considered for adoption by the International Electrotechnical Commission (IEC) and also by the U. S. Department of Defense. The system is based upon extensive work performed at the National Bureau of Standards and reported earlier.[4-6] Specifications for the necessary instrumentation were prepared by NBS, and an instrument conforming to these specifications was procured. This paper describes the system and the instrument. Comparison measurements show excellent agreement between this newer instrument and two older test sets developed at NBS and used in earlier noise studies. The design of the two earlier sets is different from that of the newer instrument, but the operating principles are the same.

The earlier test set and several duplicates were constructed at NBS using an assemblage of commercial equipment. Major modifications were necessary in adapting the commercial unit used to amplify and indicate the measured noise. All of these earlier sets were thoroughly tested and found to be completely satisfactory as to accuracy and repeatability. Recently, however, a key piece of equipment used in the set became obsolete and unavailable, making continued recommendation of this set impractical. The need for an available substitute system, based upon operating principles the same as those of the earlier set, has resulted in the present improved system. The noise quality index, conversion gain, which was originally used, has been modified. The new instrument measures current noise in terms of the modified *Index*, described in the next section.

II. Noise Quality Index

For many years a commonly used index for expressing current noise in resistors was the ratio of the mean rectified noise generated in the resistor, in microvolts, to the dc voltage applied to the resistor, in volts. A definite pass band is necessarily associated with each measurement and should always be specified, but this was often not the case. Consequently, the index did not always have the same significance in different laboratories. However, for lack of a better index, this one was used despite its disadvantages.

A more recent index, conversion gain, has been proposed[4] which does provide a definitive means for specifying noise quality. Unfortunately, it is an unfamiliar concept

* Received by the PGCP, May 4, 1960; revised manuscript received, June 20, 1960.

† National Bureau of Standards, Washington 25, D. C.
‡ Quan-Tech Labs., Inc., Boonton, N. J.

[1] D. A. Bell, "Electrical Noise," D. Van Nostrand Co., Inc., New York, N. Y., pp. 210–245; 1960.

[2] H. Nyquist, "Thermal agitation of electric charge in conductors," *Phys. Rev.*, vol. 32, pp. 110–113; July, 1928.

[3] A. von Hettich, "Spatial dimensions and resistor noise," *Frequenz*, vol. 4, pp. 14–25; January, 1950.

[4] G. T. Conrad, Jr., "A proposed current-noise index for composition resistors," IRE TRANS. ON COMPONENT PARTS, vol. CP-3, pp. 14–20; March, 1956.

[5] G. T. Conrad, Jr., "Noise measurements of composition resistors, I. The method and equipment," IRE TRANS. ON COMPONENT PARTS, vol. CP-4, pp. 61–78; November, 1955.

[6] G. T. Conrad, Jr., "Noise measurements of composition resistors, II. Characteristics and comparison of resistors," IRE TRANS. ON COMPONENTS PARTS, vol. CP-4, pp. 79–92; November, 1955.

and lacks apparent relationship to the commonly used index (microvolts-per-volt). Another shortcoming is that conversion gain is expressed in units which generally range from -140 to -200 db, magnitudes both cumbersome and difficult to visualize.

At an IEC working group meeting, a modified form of the conversion gain index, suggested by R. H. W. Burkett,[7] was agreed upon as having considerable merit. The suggested *Index* (which is italicized herein) is the ratio of the rms noise voltage, in microvolts, to the applied dc voltage, in volts, expressed in decibels, when the associated pass band for the noise is one frequency decade. This *Index*, which will be called the "microvolts-per-volt" *Index*, is definite; it bears a simple relationship to conversion gain, and it is readily associated with the more familiar and customarily used number of microvolts per volt. Because of the close relationship between conversion gain and the "microvolts-per-volt" *Index*, the same noise test system can be used to determine either index. The numerical difference between conversion gain and the "microvolts-per-volt" *Index* is a constant of approximately 160 db (to be shown below).

For the reasons given, the proposed standard noise test system described in this paper is designed to indicate current noise in terms of the "microvolts-per-volt" *Index*.

A. *Definition of "Microvolts-per-Volt" Index*

This *Index* is defined as follows:

$$Index = 20 \log \frac{v_{\text{rms}}}{V} \qquad \text{db (in a decade)}^{8}, \quad (1)$$

where v_{rms} is the *number* of microvolts of open-circuit rms current-noise voltage in a decade, and V is the *number* of dc volts applied to the resistor under test. A typical experiment might give 5 μv of noise for 50 volts of applied dc voltage. This would give a value of the *Index* of -20 db which is typical of certain groups of carbon composition resistors. The label *in a decade* is a device used to refer the *Index* to an ideal decade pass band. An ideal decade pass band is defined as a pass band having a rectangular shape with the upper cutoff frequency (f_2) equal to ten times the lower cutoff frequency (f_1). The reference ratio of the *Index*, zero decibels, has been arbitrarily chosen to be one rms microvolt per volt in a frequency decade.

It should be noted that the same amount of noise power or of mean-square noise voltage would be transmitted in a frequency decade regardless of its location in the frequency spectrum if the noise power varies inversely with the frequency throughout the spectrum of interest, *i.e.*, if the noise has an f^{-1} power spectrum. This is true not only for decade pass bands, but also for any pass bands where the cutoff frequency ratios are equal.

In general, the mean-square voltage of transmitted noise $\overline{v^2}$ is given by

$$\overline{v^2} = \int_0^\infty \frac{\overline{e^2(f)}}{\Delta f} G(f)\, df. \quad (2)$$

In particular, for an ideal pass band f_1 to f_2, (2) may be rewritten

$$\overline{v^2} = G \int_{f_1}^{f_2} \frac{\overline{e^2(f)}}{\Delta f}\, df. \quad (2a)$$

Here $\overline{e^2(f)}/\Delta f$ is the input mean-square noise-voltage spectral density at frequency f, $G(f)$ is the voltage-squared gain of the system, and G is the constant gain in the ideal pass band. A sufficiently close approximation of $\overline{e^2(f)}/\Delta f$ may be made by considering it to be the mean-square noise voltage in a one-cps pass band where the center frequency of the one-cps pass band is large compared to one cps.

Eq. (2a) may be solved by expressing the spectral density explicitly in terms of frequency. The power spectral density, which is proportional to the mean-square noise-voltage density of current noise, is commonly accepted as being proportional to f^{-1}. The effects of observed departures from f^{-1} behavior are considered in Section III. The explicit expression for the mean-square voltage spectral density is

$$\frac{\overline{e^2(f)}}{\Delta f} = \frac{C}{f}, \quad (3)$$

where C is a constant for a given noise source and represents the product of this density and frequency. We may now solve (2a) by using the fact that the product of density and frequency is a constant. Multiplying the integrand by f/f and grouping the product of density and frequency, we obtain

$$\overline{v^2} = G\left[\frac{\overline{e^2(f)}}{\Delta f} f\right] \int_{f_1}^{f_2} \frac{df}{f}, \quad (4)$$

where the density-frequency product (in brackets) preceding the integrand, even though containing the variable f, is a constant, and for this reason is not included within the integrand. Evaluating the integral and setting G equal to unity we find

$$\overline{v^2} = \frac{\overline{e^2(f)}}{\Delta f} f \ln \frac{f_2}{f_1}. \quad (4a)$$

The procedure of setting G equal to unity is a normalization procedure which simplifies interpretation of the significance of the output signal. The effect of gain is eliminated, but the effect of the system pass band is retained. This procedure is known as referring the output signal to the input.

[7] The Welwyn Electrical Labs., Ltd., Bedlington, Northumberland, Eng.

[8] The reader is cautioned to observe that both natural logarithms (ln) and common logarithms (log) are used in this paper.

If we choose f_2/f_1 to be ten then

$$\overline{v^2} = \frac{\overline{e^2(f)}}{\Delta f} f \ln 10 \qquad \text{(in a decade)}, \quad (4b)$$

where $\overline{v^2}$ is the mean-square noise voltage in a decade. For $\sqrt{\overline{v^2}}$ in microvolts, $\sqrt{\overline{v^2}} \triangleq v_{\text{rms}}$.[9] Now (1) may be rewritten

$$Index = 20 \log \frac{\sqrt{\overline{v^2}}}{V} \qquad \text{db (in a decade)} \quad (5)$$

$$Index = 20 \log \left\{ \frac{\sqrt{\frac{\overline{e^2(f)}}{\Delta f} f \ln 10}}{V} \right\} \qquad \text{db (in a decade)}. \quad (5a)$$

As previously defined, zero decibels refers to one rms microvolt (open-circuit) per volt in a decade pass band.

The dependence of the *Index* on applied dc voltage, shown in Fig. 1, illustrates experimental results[10,6] indicating that the rms current-noise voltage, v_{rms}, is proportional to the applied dc voltage raised to an exponent $\alpha/2$, with α ranging from 1.6 to 2.4. For α equal to 2, the *Index* is independent of the applied dc voltage. If the more representative value of 1.9 were used, Fig. 1 shows that for a 10 to 1 change in applied dc voltage only 1 db change in the *Index* would be expected. From this it is evident that measurement of the *Index* does not require careful selection or adjustment of the applied dc voltage. Furthermore, except for unusual resistors or operating conditions, the value of the *Index* once determined may reasonably be assumed to remain constant over a broad range of operating conditions.

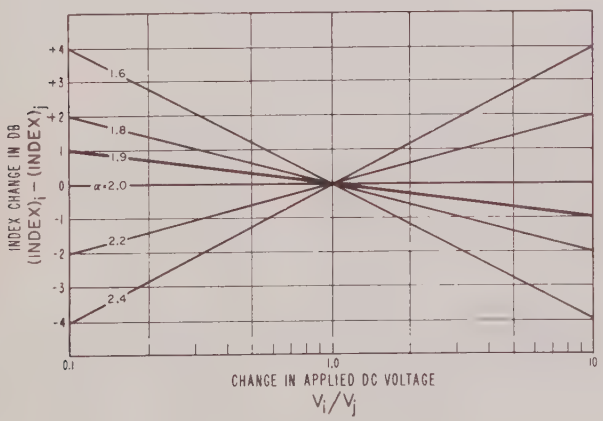


Fig. 1—Dependence of *Index* on applied dc voltage.

The "microvolts-per-volt" *Index* allows those concerned with the design of electronic equipment to estimate resistor current-noise interference in a system. In practical applications the frequency band of interest is usually not a decade. The calculation of the noise voltage in a frequency interval with any ratio of upper-cutoff frequency, f_2, to lower-cutoff frequency f_1, will now be discussed and expressed in terms of the *Index*. The current noise produced by a resistor can be described in terms of an equivalent noise-voltage generator located in series with the resistor under consideration. The rms voltage of the generator is the noise voltage which would be measured under open-circuit conditions and transmitted in the pass band f_1 to f_2. Its mean-square magnitude is given by (4a). By substituting the value for $\sqrt{[\overline{e^2(f)}/\Delta f] f}$ obtained from (5a),

$$\sqrt{\frac{\overline{e^2(f)}}{\Delta f} f} = \frac{V\, 10^{(Index)/20}}{\sqrt{\ln 10}}, \quad (6)$$

into the root expression of (4a), we get

$$\sqrt{\overline{v^2}} = V\, 10^{(Index)/20} \sqrt{\frac{\ln \frac{f_2}{f_1}}{\ln 10}} \quad (7)$$

$$\sqrt{\overline{v^2}} = V\, 10^{(Index)/20} \sqrt{\log \frac{f_2}{f_1}}$$

μv (in pass band f_1 to f_2). (7a)

The ratio from (7a) of the rms noise voltage (in microvolts) to the impressed dc voltage (in volts) expressed in db is

$$20 \log \frac{\sqrt{\overline{v^2}}}{V} = (Index) + 10 \log \log \frac{f_2}{f_1}$$

db (in pass band f_1 to f_2). (7b)[11]

As a numerical example, assume f_2 is 10 kc and f_1 is 30 cps; we find

$$10 \log \log \frac{10{,}000}{30} = 10 \log 2.52 = 4.0 \text{ db}.$$

This means that 4.0 db must be added to the value of the *Index* in order to determine the ratio (in db) of the equivalent noise generator rms voltage to the impressed dc voltage. Notice that the pass band of interest is 2.52 decades.

B. Comparison with Conversion Gain

Conversion gain is a noise quality index that was conceived and used by NBS in a series of noise studies. The relationship between the conversion gain and the "microvolts-per-volt" *Index* is given below so that the accumulation of data in terms of conversion gain may

[9] $\triangleq$ means equal by definition.

[10] C. J. Christensen and G. L. Pearson, "Resistance fluctuations in carbon microphones and other granular resistances," *Bell Sys. Tech. J.*, vol. 15, pp. 197–223; April, 1936.

[11] If the pass band is given in number of frequency decades, (7a) may be stated in the convenient form

$$\sqrt{\overline{v^2}} = V\sqrt{n}\, 10^{(Index)/20}\ \mu\text{v} \qquad \text{(in } n \text{ decades)}, \quad (7c)$$

where n is the number of frequency decades such that 10^n is equal to f_2/f_1. The form corresponding to (7b) is

$$20 \log \sqrt{\overline{v^2}}/V = (Index) + 10 \log n \qquad \text{db (in } n \text{ decades)} \quad (7d)$$

be converted to the more recent *Index*. As will be shown, the relationship between the two indexes is quite simple, and test sets designed to measure conversion gain may also be used to measure the "microvolts-per-volt" *Index*.

Conversion gain is defined as the ratio (expressed in decibels) of available mean current-noise-power spectral density, P_a, at 1000 cps to applied dc power, P_{dc}.

$$G_c = 10 \log \frac{P_a}{P_{dc}} \quad \text{db}, \quad (8)$$

or

$$G_c = 10 \log \frac{\overline{e^2}/4R}{V^2/R} \quad \text{db}, \quad (8a)$$

where $\overline{e^2}$ is measured noise in mean-square volts transmitted in a one-cycle pass band centered at 1000 cps, and V is in dc volts. Changing to the notation of this paper and expressing the noise in rms voltage, $\sqrt{\overline{v^2}}$, in microvolts, we obtain

$$G_c = 10 \log \left[\frac{\overline{v^2}}{V^2} \frac{10^{-12}}{4} \right] \quad \text{db}, \quad (8b)$$

or

$$G_c = 20 \log \left[\frac{\sqrt{\overline{v^2}}}{V} \frac{10^{-6}}{2} \right]$$

db (in a one-cps pass band at 1000 cps), (8c)

where $\sqrt{\overline{v^2}}$ is the open-circuit rms noise voltage, in microvolts, transmitted in a one-cycle pass band at 1000 cps. By expanding (8c) and using (7a) we obtain

$$G_c = (Index) + 10 \log \log \frac{f_2}{f_1} + 20 \log \frac{10^{-6}}{2} \quad \text{db}, \quad (9)$$

where a one-cycle pass band centered at 1000 cps would have essentially a frequency ratio f_2/f_1 equal to 1001/1000. Evaluating the constants we get

$$Index - G_c = 159.6 \text{ db}. \quad (10)$$

Since the indexes, as shown, differ by only a constant, it would appear that the same noise test set could be employed for measurement of either index. Small-order discrepancies would be involved for resistor current noise having departures from the f^{-1} power spectrum. Such discrepancies are negligible and are discussed more fully in Section III. Furthermore, the same measure of variability applies to either index. That is, if a group of resistors has a standard deviation of 3 db associated with the measurement of conversion gain, the same 3-db value applies to the "microvolts-per-volt" *Index*. The measured value of standard deviation, however, is a function of the variability of not only the resistor under test but also that of the test set, regardless of which of the indexes is being measured.

A measuring system for the determination of the "microvolt-per-volt" *Index* is discussed in detail in Section III.

III. Theory of Operation

A. DC Measurement Considerations

The recommended standard resistor-noise test system measures rms noise voltage referred to the terminals of an equivalent noise voltage generator located in series with the resistor under test. For determining the magnitude of current noise, two noise measurements must be made: the first when there is no dc current through the test resistor, the second when there is dc current. The first measurement indicates the "system" noise voltage which results from thermal noise of the resistor under test and all other noise sources present, such as shot noise and flicker noise of the input tubes. The second measurement, termed "total" noise voltage, represents the root of the quadratic sum of the system noise voltage and current-noise voltage, where the fluctuation voltage which causes the *increase* in noise level when dc voltage is applied is defined as current noise. It is presumed that the system noise remains constant whether or not there is dc current. The magnitude of current-noise voltage is indicated by subtracting the effects of the presence of system noise from the total noise voltage measurement. Determination of the current-noise *Index* also requires measurement of the associated value of dc voltage applied to the test resistor. Calculations involved in computing the *Index*, and the derivation of the necessary equations, are given in Section III-C.

The recommended test system for making the measurements outlined above is shown in a simplified block diagram in Fig. 2. The dc portion of the system consists of a variable dc power supply and a dc voltmeter. The ac portion of the system consists of a calibration signal source and an indicating amplifier.

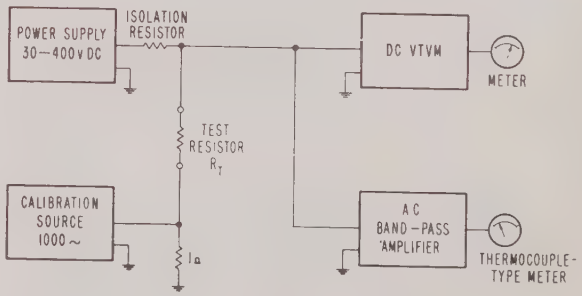


Fig. 2—Block diagram of system.

The variable dc supply furnishes dc loading power, through an isolation resistor, to the resistor under test. The isolation resistor serves to prevent the noise appearing at the terminals of the resistor under test from being severely attenuated by the very low parallel impedance presented by the output terminals of the dc supply. The isolation resistor, which must be free of current noise, could be a quiet wire-wound type. One of four values of

isolation resistor (1*K*, 10*K*, 100*K*, or 1*M* Ω) is selected depending on the resistance of the resistor under test. The recommended value of isolation resistor for a particular value of test resistor is given in Table I. Resistance values of test resistors, labeled R in this table, range from 100 Ω to 22*M* Ω in nominal 20 per cent steps. Choice of the associated isolation resistor, R_m, and the dc load voltage, V, is based primarily on three limiting factors:

1) Maximum dc power dissipated by the test resistor is limited to 250 mw.
2) Maximum dc voltage across the test resistor is limited to 250 volts.
3) Maximum dc power dissipated by the isolation resistor is limited to 1 watt.

Table I conforms to the limitations just given and to two other factors, namely, the use of only four decade values of R_m and a maximum dc supply voltage of 400 volts. These various restrictions result in the table having four distinct combinations of factors described below.

1) For values of R ranging from 100 to, but not including, 250 Ω, the dc voltage value is selected to load R_m to its maximum, 1 watt.
2) For values of R ranging from 250 Ω to, but not including, 250*K* Ω, the dc voltage value is selected to load the test resistor to 250 mw. R_m is the maximum decade value at which its power dissipation will not exceed 1 watt, the power rating of the isolation resistor.
3) For values of R ranging from 250*K* Ω to, but not including, 1.8*M* Ω, and for an R_m value of 1*M* Ω, the dc voltage value is the maximum obtainable from the 400-volt dc supply. A choice of 100*K* Ω for R_m would have permitted the selected value of dc voltage to result in 250 mw loading, but the loss in noise signal due to shunting R with the lower value of R_m would have resulted in a net loss in sensitivity in spite of the larger value of the applied dc voltage.
4) For values of R equal to or greater than 1.8*M* Ω, the dc voltage is selected to limit the maximum test voltage to 250 volts. R_m is the maximum permissible value, 1*M* Ω.

The dc power need not be adjusted to the exact value in the table. However, the value used as indicated by the dc voltmeter must be carefully read.

The dc voltage appearing across the resistor under test is indicated by the dc VTVM. The meter has two scales: one showing the voltage V, and the other indicating the quantity 20 log V, in decibels. Preferably, the meter should have a log-type movement so that the db scale is linear. The db scale simplifies computation of the current-noise *Index*, as described in detail in Section III-C.

TABLE I
RECOMMENDED OPERATING CONDITIONS

R Ω	R_m Ω	(D)* db	V volts	P_{dc} mw
100	1K	10.1	3.2	100
120	1K	11.6	3.8	120
150	1K	13.5	4.7	150
180	1K	15.1	5.7	180
220	1K	16.9	7.0	220
270	1K	18.3	8.2	250
330	1K	19.2	9.1	250
390	1K	19.9	9.9	250
470	1K	20.7	10.8	250
560	1K	21.4	11.8	250
680	1K	22.3	13.0	250
820	1K	23.1	14.3	250
1.0K	1K	24.0	15.8	250
1.2K	1K	24.8	17.3	250
1.5K	1K	25.8	19.4	250
1.8K	1K	26.6	21.2	250
2.2K	1K	27.4	23.4	250
2.7K	10K	28.3	26.0	250
3.3K	10K	29.2	28 7	250
3.9K	10K	29.9	31.2	250
4.7K	10K	30.8	34.3	250
5.6K	10K	31.5	37.4	250
6.8K	10K	32.3	41.2	250
8.2K	10K	33.2	45.3	250
10 K	10K	34.0	50 0	250
12 K	10K	34.8	54.8	250
15 K	10K	35.8	61.2	250
18 K	10K	36.6	67.1	250
22 K	10K	37.4	74.2	250
27 K	100K	38.3	82.2	250
33 K	100K	39.2	90.8	250
39 K	100K	40.0	98.7	250
47 K	100K	40.7	108	250
56 K	100K	41.5	118	250
68 K	100K	42.3	130	250
82 K	100K	43.1	143	250
100 K	100K	44.0	158	250
120 K	100K	44.8	173	250
150 K	100K	45 8	194	250
180 K	100K	46.5	212	250
220 K	100K	47.5	234	250
270 K	1M	38.6	85.0	26.8
330 K	1M	40.0	99.0	29.7
390 K	1M	41.0	112	32.2
470 K	1M	42.1	127	34.3
560 K	1M	43.1	143	36.5
680 K	1M	44.2	161	38.1
820 K	1M	45.1	180	39.5
1.0M	1M	46.0	200	40.0
1.2M	1M	46.8	218	39.6
1.5M	1M	47.6	240	38.4
1.8M	1M	48.0	250	34.7
2.2M	1M	48.0	250	28.4
2.7M	1M	48.0	250	23.2
3.3M	1M	48.0	250	18.9
3.9M	1M	48.0	250	16.0
4.7M	1M	48.0	250	13.3
5.6M	1M	48.0	250	11.2
6.8M	1M	48.0	250	9.2
8.2M	1M	48.0	250	7.6
10 M	1M	48.0	250	6.2
12 M	1M	48.0	250	5.2
15 M	1M	48.0	250	4.2
18 M	1M	48.0	250	3.5
22 M	1M	48.0	250	2.8

* $D = 20 \log V$.

B. AC Measurement Considerations

Referring to the ac portion of the system, the calibration source is a stable 1000-cps sine-wave generator. The ac indicating amplifier, as shown in Fig. 3, consists of a high-gain low-noise amplifier, a filter, and an rms detector and indicating meter. The filter restricts the frequency response of the amplifier to a flat-top 1000-cps pass band centered at 1000 cps. The indicating meter has a logarithmic response, and like the dc meter, is calibrated in decibels to simplify computations. The magnitude of the mean value of input signals may vary over a range of more than 80 db, depending upon the component being tested. To accommodate such a large range, a step attenuator is required to vary the gain of the amplifier. The magnitude of input signal is indicated by the algebraic summation of the setting of the step attenuator, calibrated in decibels, and the deflection of the indicating meter. When the system is properly calibrated, zero decibels refers to an input signal equivalent to that of a zero-impedance voltage generator located in series with the resistor under test. This signal has an rms current-noise voltage value of one microvolt in a decade.

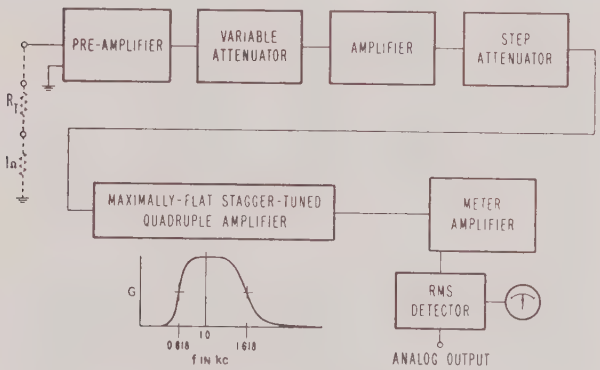


Fig. 3—Block diagram of ac indicating amplifier.

1) Calibration: Calibration consists of first applying a predetermined value of sine-wave voltage across a one-ohm resistor located in series with the resistor under test, and then adjusting the gain of the amplifier, by means of a variable attenuator, until the output meter deflects to a "calibrate" line. This procedure standardizes the gain of the system and calibrates the indicating amplifier. The resistance of the calibration resistor is selected to be negligible (1 Ω) compared to that of any test resistor (100 Ω to 22*M* Ω); therefore, the calibration voltage is equivalent to the calibration voltage appearing across a zero-impedance generator located in series with the resistor under test. The magnitude of calibration voltage is so chosen that the indicated output is equal to that which would be obtained if the calibration voltage were a noise voltage having a value of 1000 μv rms in a decade. Such a signal should produce a reading of 60 db when the system is properly calibrated.

The proper magnitude of the sine-wave calibration signal is derived by means of two different methods: 1) an analysis of an ideal measuring system having a rectangular-shaped pass band; 2) a graphical analysis of an actual system, using the data obtained from a test set described in later sections of this paper.

2) Analysis of an Ideal System: The response of a system to current noise is given in (4a), repeated here for convenience.

$$\overline{v^2} = \frac{\overline{e^2(f)}}{\Delta f} f \ln \frac{f_2}{f_1}. \tag{4a}$$

The system is considered to have an ideal rectangular pass band from f_1 to f_2 with a normalized gain of unity. Current noise is assumed to have an f^{-1} power spectrum. The term $\overline{v^2}$ represents the mean-square output noise voltage which is transmitted within the pass band f_1 to f_2. Because the gain is normalized to unity, the output signal is referred to the input. The input is considered to be the terminals of the calibration resistor, *i.e.*, the terminals of a zero-impedance equivalent generator located in series with the resistor under test. Eq. (4a) may be rearranged to solve for the associated value of density-frequency product.

$$\frac{\overline{e^2(f)}}{\Delta f} f = \frac{\overline{v^2}}{\ln \frac{f_2}{f_1}}. \tag{11}$$

Our purpose is to determine the magnitude of calibration signal which produces a 60-db meter reading. This calibration signal must produce an output voltage equal to that caused by current noise having a mean-square value of 1 mv-squared in a decade. Substituting these values of mean-square voltage and frequency ratio into (11) we obtain

$$\frac{\overline{e^2(f)}}{\Delta f} f = \frac{1}{\ln 10} \text{ mv-squared.} \tag{11a}$$

By substituting the product given in (11a) into (4a) where the pass band represented by f_1 and f_2 is the pass band of the system under consideration, we obtain the required mean-square value of the calibration signal.

$$\overline{v^2} = \frac{1}{\ln 10} \ln \frac{f_2}{f_1} \tag{12}$$

$$= \log \frac{f_2}{f_1}. \tag{12a}$$

The pass band of the system is considered to be restricted by an ideal filter having a pass band of 1000 cps geometrically centered at 1000 cps. For this filter, f_1 is 618 cps and f_2 is 1618 cps. Substituting these values in (12a) and taking the root, we obtain the required rms magni-

tude of the sine-wave calibration signal.

$$v = 0.646 \text{ mv}. \tag{12b}$$

It is appropriate to consider also the response of the system to noise having other than an f^{-1} power spectrum. For the types of noise of interest in this paper, the relationship between mean-square noise voltage spectral density and frequency (known as power spectrum) may be expressed as follows:

$$\frac{\overline{e^2(f)}}{\Delta f} = \frac{\overline{e^2(f_0)}}{\Delta f}\left(\frac{f}{f_0}\right)^{\gamma} \tag{13}$$

where $\overline{e^2(f)}/\Delta f$ is the density at any frequency f, and $\overline{e^2(f_0)}/\Delta f$ is the density at the frequency f_0. The parameter γ is a constant which takes different values for different power spectrums. For example, the thermal noise of resistors and shot noise of vacuum tubes are characterized by a "flat" or "white" power spectrum, *i.e.*, $\gamma = 0$. On the other hand, the current noise of resistors varies inversely with frequency, and hence is said to have a $1/f$ power spectrum. The corresponding value of γ is -1. Experimental results[4,12] show that the distribution of measured values of γ tend to center near the value -1 and that extreme measured values approach approximately -0.8 and -1.2. Thus, $\gamma = -1.0$ is considered representative for current noise.

At the beginning of this section the magnitude of calibration signal was derived with the assumption that $\gamma = -1$. Similar derivations for γ ranging from -0.75 to -1.25 have shown that the range of required calibration signals falls within 1 per cent of the calibration signal associated with $\gamma = -1.0$. From this we may conclude that the proposed system provides a direct measurement of current noise independent of the value of γ throughout the entire range from -0.8 to -1.2 listed in the reports referenced.[4,12] The measurement provides an *Index* of current noise within one frequency decade. The magnitude of noise in a decade is independent of the location of the decade in the frequency spectrum only for the case where $\gamma = -1$. Otherwise the location is a factor, and although as a practical matter it is relatively small, it should be specified. The reference decade pass band of the proposed system is centered at 1000 cps.

It has been shown that variations in γ do not affect the measurement of the current-noise *Index*. However, when a resistor is measured, the measuring system provides no indication of the value of γ associated with a particular resistor measurement. One may therefore be concerned as to the error that may be involved in estimating interference on the basis that $\gamma = -1$ when the actual value of γ is not known. An estimate of the limit of such an error may be made by substituting one of the extreme values of γ, say $\gamma = -0.8$, into (14), which holds for the range $-0.8 \geq \gamma \geq -1.2$.

$$\frac{\sqrt{\overline{v^2}}}{\sqrt{\overline{v^2}}\,|_{\gamma=-1}} = \left[\frac{\sqrt{f_1 f_2}}{1000}\right]^{(\gamma+1)/2} \tag{14}$$

Eq. (14) is a correction factor multiplier for (7a) which may be used to provide a better estimate of noise transmitted in the pass band f_1 to f_2 when γ is known and falls within the range -0.8 to -1.2. This correction factor is an approximation which stems from the assumption that the mean-square noise voltage transmitted in a pass band varies with the density at the center frequency of the pass band. The approximation given by (14) is valid within the range 10 cps to 100 kc. An example of the maximum error which may be expected for a typical audio pass band of 20 cps to 20 kc and $\gamma = -0.8$ is obtained by substituting these values in (14). We obtain the value 0.95 which means that the rms noise voltage transmitted would be approximately 5 per cent (0.4 db) less than that calculated on the basis of $\gamma = -1$ using (7a). The magnitude of this error is negligible. As would be expected, errors associated with a pass band having a center frequency well removed from 1 kc would be larger. For example, consider again the maximum error associated with $\gamma = -0.8$, for the pass band 1 kc to 100 kc. This would result in the solution of (14) being 1.6: *i.e.*, the transmitted rms noise voltage would be 60 per cent (4 db) greater than that predicted on the basis $\gamma = -1$.

Rarely do resistors have γ values approaching the extreme values used in the examples given above. For this reason, (7a) does not require the use of the correction factor given in (14). Also, the effects of variations in γ, as reflected by the correction factor, are relatively small.

The noise measuring system may also be used to measure noise having a "flat" power spectrum. The conditions for which the response of the system is the same for both f^0 and f^{-1} noise are obtained by substituting (13) into (2a) and then setting the solution to (2a) for $\gamma = -1$ equal to that for $\gamma = 0$.

$$\frac{\overline{e^2(f)}}{\Delta f} f \ln \frac{f_2}{f_1} = \frac{\overline{e_0^2(f)}}{\Delta f}(f_2 - f_1). \tag{15}$$

Note that the density of f^0 noise given by $\overline{e_0^2(f)}/\Delta f$ equals that of f^{-1} noise given by $\overline{e^2(f)}/\Delta f$ at the frequency

$$f = \frac{f_2 - f_1}{\ln \frac{f_2}{f_1}}. \tag{16}$$

Recalling that $f_2 = 1618$ cps and $f_1 = 618$ cps,

$$f = 1038 \text{ cps}. \tag{16a}$$

The f^0 density which would produce the same deflection as the calibration signal (60-db reading) may be calculated from (11) and (11a) where the value 1038

[12] P. L. Kirby and G. P. Sibilia, "An investigation of current noise in fixed resistors," *Proc. Symp. on Noise in Fixed Resistors*, Radar Research Establishment, Malvern, Eng., pp. 1–25; March, 1959.

cps is substituted for f. The resulting solution is

$$\frac{\overline{e^2(f)}}{\Delta f} = \frac{1}{1038 \ln 10}$$

$$= 420 \text{ mv-squared per cps.} \quad (17)$$

If we wish to have zero db represent one microvolt-squared per cps, then the above number is 26.2 db greater than the reference. Hence, the meter reading of 60 db represents the measured density value 26.2 db, and therefore the difference, 33.8 db, is the number of db which must be subtracted from the meter reading to determine the density of "flat" noise. This procedure also gives the density of current noise at the frequency 1038 cps, with the same zero db reference.

3) Analysis of an Actual System: A test set has been constructed in accordance with requirements of the noise measurement system described in this paper. A procedure for determining the response of the actual system to noise and determining the required magnitude of the calibration signal is outlined below. This procedure parallels the analysis of an ideal system given in the preceding section.

The response of a system to noise is given in (2). By substituting (3) in (2), we obtain the response to current noise which has an f^{-1} power spectrum.

$$\overline{v^2} = C \int_0^\infty \frac{G(f)}{f}\, df \quad (18)$$

where C is a constant for a particular current-noise source and is equal to the density-frequency product.

$$\overline{v^2} = \frac{\overline{e^2(f)}}{\Delta f} f \int_0^\infty \frac{G(f)}{f}\, df. \quad (18a)$$

This equation may be used to determine the response of an actual system to noise and to determine the required magnitude of calibration signal, provided that the proper constraints are satisfied in solving (18a). First, the value of $\overline{v^2}$, which represents the mean-square output voltage, must be referred to the input by normalizing the gain of the system. This is done by dividing the gain function $G(f)$ by its maximum value G. Second, the density-frequency product should represent noise which would produce a mean-square output voltage of one mv-squared in a decade (60 db). Eq. (11a) gives the proper value for the density-frequency product. When the above two restrictions are substituted in (18a), we obtain

$$\overline{v^2} = \frac{1}{\ln 10} \int_0^\infty \frac{G(f)}{G} \frac{df}{f}, \quad (18b)$$

which equals the required magnitude of the calibration signal.

Expressed as an rms voltage value and substituting A for the integral, (18b) becomes

$$v = \sqrt{\frac{1}{\ln 10} A}. \quad (18c)$$

The term A represents the graphical evaluation of the integral in (18b). The value of A is determined by first measuring and plotting the normalized gain of the system vs frequency on linear graph paper. The gain is the square of the ratio of the output voltage to the input voltage, where the input is the voltage applied across the terminals of the calibration resistor.[13] Next, the normalized gain values of the various points of the plot are divided by their corresponding frequency values and the resulting values are plotted against frequency. The area under this curve equals A. For the particular test set under consideration, A was found to be 0.968. Hence, the evaluation of (18c) for the required calibration voltage is

$$v = 0.648 \text{ mv.} \quad (18d)$$

This compares very well with the value 0.646 mv given in (12b) for the ideal system.

It should be recalled that the response of the ideal system is the same for f^0 or f^{-1} noise spectrum when the density values of both are equal at the frequency f derived in (16). The corresponding frequency f may be readily derived for the measuring set in question. The numerator of (16), $f_2 - f_1$, is replaced by its corresponding value obtained in the graphical solution for the effective pass band (1002). At the same time, the denominator of (16), $\ln (f_2/f_1)$, is replaced by its corresponding value obtained in the graphical evaluation of A, (0.968). These substitutions give

$$f = \frac{1002}{0.968} \quad (19)$$

$$f = 1036 \text{ cps.} \quad (19a)$$

Here again the value obtained from graphical analysis of an actual system agrees well with that obtained from theoretical considerations of an ideal system, where $f = 1038$ cps, as given in (16a).

C. Computations

Computation of the "microvolts-per-volt" *Index* requires measurement of three quantities: system noise, total noise, and applied dc voltage. As will be shown, necessary computations are simplified if the three measurements are made in decibel units. For the two noise measurements, zero db refers to one microvolt in a decade. For the dc voltage measurement, zero db refers to one volt.

The symbol S represents a measure of the system noise when there is no dc current through the test resistor. System noise is the quadratic sum of various extraneous noise sources, such as thermal noise of resistors and noise

[13] The area under the resulting curve is equal to the integral of the normalized gain and is equal to the effective pass band (for "flat" noise). For the particular measuring set in question, the effective pass band was found to be 1002 cps. This compares well with the 1000-cps value of the ideal system.

generated in the input tubes and transistors, *i.e.*, flicker noise, shot noise, etc.

The symbol T represents a measure of the total noise present when there is dc current. All of the system noise sources mentioned are present at the same time and are presumed to remain unchanged. The increase in thermal noise of the test resistor due to dc current is considered negligible. The total noise is greater than system noise; the increased noise is caused by current noise generated in the resistor under test. The magnitude of current noise may be calculated by taking the difference between the measured mean-square voltages; that is,

$$\sqrt{\overline{v^2}} = \sqrt{\overline{v_T^2} - \overline{v_S^2}}, \tag{20}$$

where $\sqrt{\overline{v^2}}$ is rms voltage of the current noise generated in microvolts in a decade, v_{rms}. The other voltages are those, referred to the input, which produce the decibel readings T and S.

$$\overline{v_T^2} = 10^{T/10} \tag{21}$$

$$\overline{v_S^2} = 10^{S/10} \tag{22}$$

By substituting these values for $\overline{v_T^2}$ and $\overline{v_S^2}$ in (20) and then replacing v_{rms} of (1) by the right side of (20), we obtain the expression for the *Index* in terms of the measured quantities.

$$Index = 20 \log \frac{\sqrt{\overline{v_T^2} - \overline{v_S^2}}}{V} \tag{23}$$

$$Index = 20 \log [10^{T/10} - 10^{S/10}]^{-1/2} - 20 \log V \tag{23a}$$

or

$$Index = T - f(T-S) - D \tag{23b}$$

where

$$D = 20 \log V \tag{24}$$

and

$$f(T-S) = -10 \log [1 - 10^{-(T-S)/10}]. \tag{25}$$

The function $f(T-S)$ is an additive factor which corrects for the presence of system noise while T is being measured. A tabulation of $f(T-S)$ vs $T-S$ is given in Table II. The value of $f(T-S)$ is zero and may be ignored whenever $T - S$ is greater than 15 db.

TABLE II
Correction Factor for Presence of "System Noise"

$T-S$ db	$f(T-S)$ Correction Factor	$T-S$ db	$f(T-S)$ Correction Factor	$T-S$ db	$f(T-S)$ Correction Factor
0.2	13.5	3.5	2.5	7.0	
0.3	11.8	3.6	2.4	to	0.9
0.4	10.5	3.7	2.4	7.3	
0.5	9.6	3.8	2.3	7.4	
0.6	8.9	3.9	2.2	to	0.8
0.7	8.3	4.0	2.2	7.9	
0.8	7.7	4.1	2.1	8.0	
0.9	7.3	4.2	2.0	to	0.7
1.0	6.9	4.3	2.0	8.5	
1.1	6.5	4.4	1.9	8.6	
1.2	6.2	4.5	1.9	to	0.6
1.3	5.9	4.6	1.8	9.3	
1.4	5.6	4.7	1.8	9.4	
1.5	5.3	4.8	1.7	to	0.5
1.6	5.1	4.9	1.7	9.9	
1.7	4.9	5.0	1.6	10.0	
1.8	4.7	5.1	1.6	to	0.4
1.9	4.5	5.2	1.5	11.5	
2.0	4.3	5.3	1.5	11.6	
2.1	4.1	5.4	1.4	to	0.3
2.2	3.9	5.5	1.4	12.7	
2.3	3.8	5.6	1.4	12.8	
2.4	3.6	5.7	1.3	to	0.2
2.5	3.5	5.8	1.3	14.5	
2.6	3.4	5.9	1.3	14.6	
2.7	3.3	6.0	1.2	to	0.1
2.8	3.2	6.1	1.2	15.0	
2.9	3.1	6.2	1.2		
3.0	3.0	6.3	1.1		
3.1	2.9	6.4	1.1		
3.2	2.8	6.5			
3.3	2.7	to	1.0		
3.4	2.6	6.9			

IV. Description of Test Set

Here and in Section V a test set is described which was built to conform with requirements of the recommended test system and which has proven to operate satisfactorily.

The test set is complete and self-contained in a single case measuring approximately 12 × 12 × 16 inches. The power source requirement is a 117-volt, 50- to 60-cps power line. Certain extra features, such as analog-voltage output jacks and means for identifying meter-multiplier switch positions by coded contact closures, provide the nucleus for automating measurements.

The test set is divided into two separate measuring systems. One system measures fluctuation voltage and the other measures dc voltage applied to the resistor under test. The description that follows has also been divided into two parts, the ac measuring system and the dc measuring system. A photograph of the test set is shown in Fig. 4.

A. AC Measuring System

1) Range:

a) Resistance: Resistance values ranging from 100 ohms to 22M ohms may be tested in the test set.

b) Noise: The test set is capable of measuring noise having a density from approximately $10^{-2}\mu v/\sqrt{\text{cps}}$ to $20\mu v/\sqrt{\text{cps}}$ for noise having either an f^0 power spectrum or an f^{-1} power spectrum. For the latter case the values refer to the noise spectral density at approximately 1 kc.

The range of density given above may be specified in decibel terms for rms microvolts in a decade by the equivalent range, −6.4 to +59.6 db in a decade, where 0 db refers to 1 μv in a decade. For accommodation of differences in system noise for different test sets, the scale range extends from −20 db to +60 db.

2) Input:

a) Impedance: The impedance of the amplifier measured at the test resistor terminals is relatively high, having a value of 1 megohm resistance shunted by 15 picofarads (10^{-12} farads).

b) Noise: When the test resistor terminals are shorted, the indicated noise, referred to the input, does not exceed 10^{-2} μv/$\sqrt{\text{cps}}$ at 1 kc, or −6.4 db in a decade. This is equivalent to the thermal noise of a 6*K* ohm resistor.

c) Auxiliary features: An additional receptacle is provided in the rear of the set as an auxiliary input. The auxiliary input allows operation of the test set in a normal manner, with the advantage that connections may be made to a remote pair of test-resistor terminals through a cable that may be as long as ten feet and has a driven shield for nullifying the effects of cable capacitance.

3) Attenuator: The test set contains a four-position (ac) attenuator which when stepped clockwise increases attenuation in 20 db steps.

Additional wafer-switch sections providing 4 poles and 4 positions, and having break-before-make contacts, are mounted on the attenuator shaft. These extra switches may be wired by the user to an output receptacle [see Section IV-A, 6)] to provide coded shaft position information. The choice of code is left to the user because of the wide variety of available supplementary equipment.

4) Pass Band: The pass band of the amplifying system, as measured at the output with an input signal applied across the 1-ohm calibration resistor remains constant to within ±3 per cent for all combinations of test resistors and isolating resistors. The test set pass band is determined by the filter properties of a four-section stagger-tuned amplifier with a Butterworth-shape frequency response having a half-power pass band of 1000 cps geometrically centered at 1000 cps.

5) Detector: A true rms-type detector is used to drive the noise indicating meter and to provide a dc analog output voltage made available from a jack.

6) Output:

a) Meter: The scale range is from −2 db to +20 db with approximately linear spacing. The scale is marked in 1-db intervals with every fifth number from 0 through 20 enumerated. Scale length is approximately 3½ inches. Instantaneous peak response is linear to at least +30 db, with average power delivered to the detector limited to a safe value. The time constant when noise is being measured is approximately 1 second. Accuracy is within ±2 per cent of the voltage reading (0.2 db). The meter face indicates "rms noise voltage" and "0 db = 1 μv in a decade." A thin red line at 10 db sets off the standard calibration deflection.

b) DC noise-analog output-voltage jack: The range of voltage available from this jack is from 1 to 10 volts, corresponding to ac input voltage amplitude which produces ac meter readings from 0 to 20 db. Instantaneous response and time constants are the same as those specified for the output meter. The driving resistance is not more than 1000 ohms.

c) AC output-voltage jack: The amplified and filtered ac output voltage is available at this jack. Approximately 1 volt rms corresponds to a meter reading of 20 db. The driving resistance is not more than 100 ohms.

d) Attenuator-position output receptacle: An auxiliary 36-pin connector mounted on the rear of the chassis can be wired to the position-coded wafers of both the ac meter-range and the dc meter-range attenuators in any manner required by the user's recording equipment.

7) Test Resistor Range: The test resistor range is selected by a four position rotary switch whose function is to insert one of four different decade values of current-noiseless (wire-wound), 1-watt isolating resistors, R_m, in series with the dc supply. The value of the isolating resistors and associated range of test resistors is as follows:

Range (*K*Ω)	R_m (*K*Ω)
0.1 − < 2.5	1
2.5 − < 25	10
25 − < 250	100
250 − < 22*M*Ω	1*M*Ω

8) Function Switch: The function switch is a three-position rotary switch which selects the mode of operation: calibration, the measurement of system noise, and the measurement of total noise.

a) Calibration: When the function switch is in calibration position, the gain of the system is standardized by adjusting the "calibration adjust" gain control so that the meter deflects to a red line at the 10-db point on the meter face. When this is done, the amplifier gain is standardized so that a zero-db noise signal appearing across the calibration resistor (1 ohm) would produce an output reading (sum of meter and attenuator readings) of zero db if no other signal sources were present.

The internal calibration signal source is a stable 1000-cps generator. Its output, properly attenuated, is applied across a one-ohm calibration resistor which is located between the "low" terminal of the test resistor jig and ground. The magnitude of calibration voltage is adjustable to a value calculated to produce an output meter reading equivalent to that produced by a noise generator having an rms voltage of 1000 μv in a decade. Such a signal should produce a reading of 60 db. The proper calibration signal magnitude appearing across the one-ohm resistor

was derived in Section III and found to be 0.65 mv rms. This signal should be more than 20 db larger than the sum, referred to the input, of all noise sources present during operation when dc current is not flowing through the test resistor.

Concurrent with application of the calibration signal, a meter-multiplier attenuator which would "add" 50 db to the meter reading is switched into the amplifier circuit replacing the meter-multiplier step attenuator. The sum of the attenuator (50 db) and meter deflection (10 db) equals the magnitude of the calibration signal (60 db).

The calibration gain control permits the gain of the amplifier to be varied over a range of at least 33 db. This allows the system to be calibrated with the same magnitude of calibration signal regardless of the input-circuit attenuation produced by the range of combinations of test and isolation resistors. Calibration signal amplitude accuracy is within ±0.5 per cent.

A calibration input jack mounted in the rear allows an external signal generator having a 600-ohm driving resistance and 10-volt open-circuit output to replace the internal generator. A 1000-ohm series resistor permits the calibration voltage to be determined by a measurement at the external signal generator terminals. Pass-band measurements may thus be conveniently made using an external generator.

b) System noise: The function switch in the "system noise" position allows the measurement of the test set system noise, which includes the thermal noise of the resistors in the input.

c) Total noise: The function switch in the "total noise" position differs from the "system noise" position only in that dc voltage is applied across the resistor under test, through the isolating resistor.

9) Accuracy of Noise Measurement: The component of mean-square noise voltage appearing in the output and generated in circuits following the attenuator should not exceed 2 per cent, or 0.1 db, of the total mean-square noise voltage present at the output under all normal operating conditions where the output meter deflection is within the range −2 to +20 db. To satisfy this condition a change of 20 db in the attenuator setting should change the output meter reading at least 19.9 db within its scale range insofar as noise signals are concerned, neglecting both attenuator and meter-tracking errors. Furthermore, when the output meter deflection is at its lowest reading, −2 db, and when the attenuator is stepped to introduce 20 db of attenuation, the dc analog (and ac) output voltage should decrease at least 15 db

Accuracy of noise voltage measurements is within ± 5 per cent (0.4 db).

B. DC Measuring System

1) Voltage Supply: The dc voltage supply is adjustable from 30 to 400 volts with 1 per cent regulation. This permits the test voltage appearing across the test resistors to be adjustable from approximately 3 to 300 volts. The power supply should be sufficiently hum-free that no observable increase appears in the "system noise" reading when dc power is applied to a wire-wound resistor inserted in the test jig.

2) Voltmeter: The voltmeter, which is connected permanently to the "high" terminal of the test resistor jig, has an input circuit consisting of a series-connected, current-noiseless (wire-wound) resistor having at least $4M\ \Omega$ resistance. The meter-side terminal of the resistor is shunted to ground with a capacitor having a value sufficiently large to prevent noise generated in the dc voltmeter from affecting the current-noise measurement. The meter ranges from 1 to 10 volts and has a logarithmic response. The scale length and meter are the same as those for the ac output meter. Two scales are required: one to indicate voltage, V, from 1 to 10 volts, and a corresponding second scale to indicate $20 \log V$, from 0 db to 20 db. The db scale is marked at 1-db intervals with values enumerated at 0, 5, 10, 15, and 20 db. A second scale, in volts, is marked at integral values 1 through 10 and enumerated at 1, 2, 5, and 10.

3) Attenuator: A three-position rotary switch is included for extending dc meter readings as follows:

Position	*Voltage Multiplier*	*DB "Multiplier"*
CCW	1	0
Center	10	20
CW	100	40

The combined accuracy of the voltmeter and attenuator is within ±3 per cent.

The dc attenuator shaft has an additional wafer switch identical to the one described for the ac attenuator except that only 3 positions are required.

4) Output:

a) DC analog output-voltage jack: The range of voltage available from this jack is from 1 to 10 volts corresponding to the dc voltage amplitude which produces a meter reading of from 1 to 10 volts. The driving resistance is not more than 1000 Ω.

C. Mechanical Layout

The items that appear on the front panel of the instrument are shown in Fig. 4. The following additional items appear at the rear:

Auxiliary input jack
DC noise-analog output-voltage jack
AC output noise-voltage jack
DC analog voltage jack
Calibration input-voltage jack
36-pin receptacle
Fuse
Power cord

V. Distinctive Design and Construction Features

Design and construction features of certain portions of the test set are peculiar to this instrument. The discussion which follows emphasizes description of critical circuits and critical construction features. It also includes a review of uncommon features which will clarify understanding of the instrument's operation. Descriptions of familiar circuits, such as linear amplifiers, voltmeters, power supplies, attenuators, and other conventionally designed circuits are not included.

It is intended that the instrument serve both as a laboratory measurement standard and as a production-type testing device capable of semi-automatic operation. Accordingly, in the design of the test set prime consideration was given to reliability and ease of operation. The combined use of transistors and vacuum tubes has resulted in superior operating performance and considerable reduction in size and power consumption.

A. DC Circuits

A basic block diagram of the test set is shown in Fig. 2. The source of dc voltage applied to the test resistor when current noise is being measured is a well-regulated variable dc supply essentially free of hum and noise. The presence of unwanted hum and noise would add to the indicated current noise and would be especially troublesome when measuring resistors having relatively little current noise.

The isolation resistors, which prevent the low output impedance of the dc supply from effectively shorting the terminals of the test resistor, must be carefully selected and located. These resistors, four in all, must not generate current noise or introduce hum or other interference from extraneous coupling. The generation of current noise would add to the indicated current noise. The interference from extraneous coupling would be present whether dc current flowed or not, and it would therefore simply reduce the sensitivity of the set. Interference from hum is more troublesome than that from random noise, because hum is a function of many variables, such as location of both components and wiring, and choice of both shielding and grounding system. Hum interference is particularly insidious when two (or more) such signals are present, because the observed combined effect of these signals is dependent upon their phase relationship as well as their amplitudes.

The dc voltage applied across the test resistor is continuously monitored by a VTVM, which has its indicating meter mounted on the front panel. Connecting the VTVM to the test resistor through a low-pass RC filter prevents the VTVM from introducing noise into the noise-measuring circuitry.

B. Calibration Circuits

The ac portion of the test set is comprised of a calibration source and an ac indicating amplifier. The calibration source is a transistor oscillator which generates a 1000-cps sine wave having a well-regulated output. The accuracy of the noise measurement can be expected to be no greater than the accuracy of the calibration signal. Because the calibration signal variations must be minimized, the signal amplitude is regulated to within 0.5 per cent. This, in turn, requires that the dc supply for the oscillator be well-regulated. Also, distortion must be minimized so that the rms value of the calibration signal may be accurately determined by commonly available instruments having linear detectors. This is accomplished by using a high-Q oscillator tank circuit and operating it at a low level. The frequency and stability of the calibration signal are not critical because of the broad pass band of the test set. The oscillator output is connected directly across the 1-ohm resistor shown in series with the test resistor. The effect of the calibration technique adopted is to standardize the response of the system to a voltage appearing across an effectively zero-impedance generator located in series with the resistor under test.

An input jack has been provided for test purposes and is useful for monitoring the operating characteristics discussed above. The jack permits connection of an external generator, through a 1000-ohm series resistor, to the 1-ohm calibration resistor. When a plug is inserted into the jack, a mechanical interlock turns off the calibration signal by removing bias power supplied to the oscillator.

C. Indicating Amplifier

The ac indicating amplifier is the heart of the noise-measuring system. Although the design principles used are neither new nor unique, considerable care must be taken in the design and construction of this amplifier. An expanded block diagram is shown in Fig. 3. Careful attention must be given to operation of the subsections as well as to the integrated operation of the whole. This involves consideration of signal levels, noise levels, dynamic range requirements, isolation requirements, and susceptibility to hum and other sources of interference. The two more critical subsections are the preamplifier, and the stagger-tuned amplifier which determines the band-pass characteristics of the set. These are described in the following sections.

1) Input: The input circuit is extremely sensitive; however, operational tests have shown that the simple shielded enclosure for housing the test resistor, as shown in Fig. 4, is adequate for elimination of extraneous coupling. The test resistor need only be dropped into place between the V-shaped slots of the hardened silver-alloy terminals and the shield closed before a measurement is made.

The preamplifier consists of a cathode follower coupled to a transistor amplifier. The schematic diagram is shown in Fig. 5. The hybrid vacuum tube and transistor circuit was chosen because it provides an input circuit having relatively low noise combined with a high impedance.

Fig. 4—Front view of resistor-noise test set.

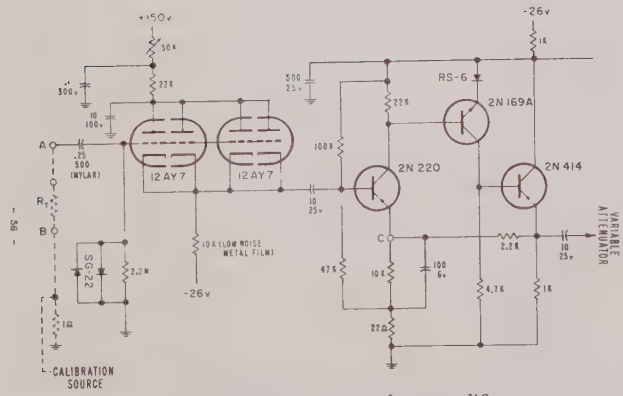


Fig. 5—Schematic diagram of preamplifier.

The cathode follower consisting of four triode sections in parallel increases the effective transconductance so as to reduce the effect of shot noise. Adjustment of the plate voltage to approximately 50 volts gives minimum noise while retaining gain stability. The voltage adjustment control is conveniently located at the rear of the instrument.

For noise measurements on resistors at a remote location from the test set, a triple shielded cable with one shield driven can be conveniently used for interconnection. Fig. 5 shows the instrument input circuit. The center conductor of the cable connects point A with the high side of the test resistor at its remote location. The innermost shield, the driven shield, connects to point C. It has no remote connection. The second shield connects point B to the low side of the test resistor. The outermost shield connects the remote ground to the instrument ground.

The transistor portion of the preamplifier has an amplification of approximately 100. Magnetic shielding of the preamplifier, the variable attenuator, and the following amplifier is essential. Special precautions should be taken in the layout and selection of ground tie points to minimize the effects of inductive pickup loops.

2) Filter: The band-pass characteristics of the test set are established by a maximally-flat stagger-tuned quad-

ruple amplifier which follows the meter-multiplier step attenuator. Since the signal levels of the attenuator output are relatively low, the amplifier should have sufficiently low noise to prevent the sensitivity of the test set from being appreciably degraded. Each of the four tuned sections (see Fig. 6) consists of a transistor amplifier followed by a series-tuned LC circuit, resistance loaded to provide the proper Q. The amplifier portion of each section consists of a three-transistor amplifier stabilized by heavy feedback and having an emitter-follower output to insure a low driving impedance. Each amplifier has an input resistance of approximately $100K$ ohms and a driving impedance of less than 5 ohms. Each serves primarily as a buffer and incidentally provides an amplification of approximately 10.

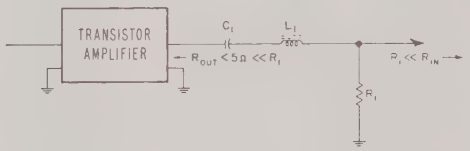


Fig. 6—Diagram of filter section F_1.

The tuned-circuit constants may be determined from the maximally-flat band-pass equations described by Butterworth. The geometric center frequency of the desired pass band is 1000 cps and the half-power pass band is 1000 cps. The resulting filter skirts are sufficiently sharp to attenuate effectively interference of power-line frequencies and their more troublesome lower harmonics. Suitable values of L, C, and R for each filter section may be estimated from the filter constants listed in Table III. In the alignment of each filter section, the inductance and/or capacitance is tuned to the frequency listed above. Then the terminating resistor is adjusted to provide the proper Q.

TABLE III
Filter Constants

Filter Section	Center Frequency in kc	Q
F_1	0.635	2.95
F_2	0.778	1.158
F_3	1.283	1.158
F_4	1.575	2.95

3) Detector: The meter amplifier, which follows the final filter section, is channeled to two outputs. A true rms detector, described below, is used to obtain these outputs. One output provides a dc voltage which is the analog of the rms value of the noise voltage being measured. The other output drives the front-panel meter which indicates, on a db scale, the rms value of the noise being measured. The indicating meter has a suppressed zero as well as shaped pole pieces to provide a linearly spaced db scale marked from −2 to 20 db. The dynamic linear range of the meter amplifier extends at least 12 db beyond the full-scale meter reading for instantaneous peak signals. This additional range is essential for accurate measurement of signals having the relatively large peaks of short duration, characteristic of noise.

A true rms detector is capable of measuring the rms value of a fluctuating voltage (or current) regardless of amplitude distribution or wave form. In certain cases, a linear detector may be used for measuring rms values, but the constant of proportionality between mean rectified voltage and rms voltage must be known for the particular amplitude distribution of the signal being measured. The constant is known for noise signals having a Gaussian amplitude distribution which is an observed characteristic[4] of current noise. The true rms detector used in the test set enables the set to indicate rms values regardless of amplitude distribution. For this reason, the test set may serve as a general-purpose noise-measuring instrument having the ability to measure non-Gaussian as well as Gaussian type noise.

One of the simplest rms detectors which would be suitable is a thermocouple meter, but its effective indicating range would ordinarily be limited to less than 20 db. Use of a thermocouple meter would therefore require changing the meter multiplier attenuator from four 20-db steps to seven 10-db steps. Also, additional circuitry would be needed for furnishing a dc analog output voltage. The rms detector used in the test set differs from the thermocouple meter type and has certain advantages as will be described.

The test set rms detector system, which is necessarily of somewhat sophisticated design, provides a dc current (and voltage) proportional to the rms value of the input signal voltage over a 22-db range. The operating principle is similar to that employed in typical AVC systems of radio receivers. A measure of the amplitude of an incoming radio signal, which may vary over a wide range, is obtained by monitoring the AVC dc-feedback voltage with an "s" meter. This dc-feedback voltage serves to reduce the receiver sensitivity with increasing values of the incoming signal. The receiver sensitivity is controlled by a variable-gain amplifier whose amplification is reduced by increasing the magnitude of the dc feedback voltage, and vice versa. By a rectification and averaging process, the dc feedback voltage is derived from the signal that is transmitted by the variable-gain (compression type) amplifier. The "s" meter may be calibrated to indicate the magnitude of the incoming radio signal.

Similarly for the test set, the measurement of the rms value of an input signal voltage, whose rms value may fall within a 22-db range, is obtained by monitoring the dc-feedback current with the front-panel noise meter. This current feedback serves to reduce the voltage amplification of the compression amplifier directly with in-

creasing values of the rms input noise voltage, and vice versa. The monitored dc-feedback current, which indicates the rms value of the input signal, is obtained by use of a hot-wire thermal element as the detector. The thermal element is a 26-volt incandescent lamp connected across the output of the compression amplifier. The dc-feedback current to the compression amplifier is derived from a circuit which monitors the resistance of the thermal element. The constants of this same circuit are so adjusted that the feedback current is proportional to the rms value of the input noise voltage within the operating range of the detector. It is this feedback current which is monitored with the front-panel noise meter.

For a range of 10 to 1 (20 db) of the rms value of the input voltage, and the same range of the feedback current, the action of the compression amplifier provides a corresponding voltage range of less than 1.1 to 1 at its output terminals. The hot-wire thermal element, which is driven by the compression amplifier, is therefore restricted to an operating voltage range of less than 1.1 to 1. This is desirable for several reasons. A limited operating range tends to minimize second- and higher-order errors associated with the quadratic detector. The inherent time constant of the smaller operating range of the detector is reduced to a value of approximately 0.1 second; hence, the time constant of the measuring system may be increased to any desired value with RC elements in the dc feedback circuit. Furthermore, since the operating temperature of the thermal element has been restricted to a narrow operating range and has been conservatively chosen, it is much lower at all times than its burnout temperature. The amplifiers in the set will limit, well before the burnout temperature of the thermal element is reached.

The noise voltmeter, like the dc voltmeter, has an associated analog output which provides 10 volts dc, corresponding to a full-scale noise meter deflection of 20 db. This output may be used for remote monitoring or recording of the dc voltage analog of the noise. Means are also available for monitoring or recording the meter-multiplier attenuator position. Such means are in the form of unwired wafer switches on the shaft of the attenuator. A similar arrangement is also provided for the dc measuring meter. These various outputs may be used to operate external control circuits for automating go-no-go or other types of measurements.

D. Stability

It is emphasized that the pass band of the system must remain stable. This dictates the use of stable, high-quality components. Also, the calibration signal amplitude must be stable. Its frequency stability is not particularly critical because the pass band is flat in the frequency neighborhood of the calibration signal. For normal operation, the test set gain stability is not critical because the calibration procedure compensates for gain drift.

VI. Test Set Operating Procedures

Before a test set can be used for making noise measurements, the proper magnitude of calibration signal must be determined and the calibration generator output voltage set accordingly. The procedure for doing this is outlined in Section III-B, 3). The necessary gain measurements may be made merely by substituting an external audio oscillator for the internal calibration generator by way of the calibration input jack. The resulting plot of *normalized* gain vs frequency must be constant and independent of the resistance value of the resistor placed across the test-resistor terminals while gain measurements are being made. This relationship should be checked for the specified extreme values, 100 ohms and 22M ohms. Once the required magnitude of calibration signal is established, the internal calibration generator may be set by means of an internal potentiometer adjustment. The set is then ready for operation.

Operation consists of three steps: 1) calibration, 2) measurement of system noise, and 3) the simultaneous measurement of applied dc voltage and the resulting "total" noise. Generally, the measurements should be made in the order listed. A detailed description of each of these steps follows.

1) Calibration. After the resistor to be tested is inserted in the test jig, the desired isolation resistor is selected by setting the RESISTOR RANGE switch to the proper range as indicated on the front panel. The system is now ready for the first step, calibration. The FUNCTION switch is set on "calibrate" and the CALIBRATE ADJUST control is varied so that the noise-voltage meter deflects to the calibration red line. The position of the METER-MULTIPLIER switch is immaterial because the FUNCTION switch, when set in the calibrate position, effectively removes the METER-MULTIPLIER attenuator from the circuit and inserts a separate suitable attenuator step.

2) Measurement of system noise. After calibration, the FUNCTION switch is set to its second position, SYSTEM NOISE (S), and the METER-MULTIPLIER switch is adjusted to make the meter read on scale. The algebraic sum of the db readings of the METER-MULTIPLIER position and the meter deflection is the measurement of "system" noise, S, in db.

3) Simultaneous measurement of applied dc voltage and the resulting "total" noise. The proper value of dc voltage is applied and then it and the resulting "total" noise are measured. When this is done one should *first* be certain that the dc VOLTAGE ADJUST control is set to minimum before the FUNCTION switch is set to its third position, TOTAL

NOISE (T). The recommended value of dc voltage, given in Table I, is set by the dc VOLTAGE ADJUST control. The sum of the db readings of the dc METER-MULTIPLIER attenuator position and dc meter deflection equals the db value, D, of the applied dc voltage. The associated noise measurement indicates the total noise T, in db.

The current noise *Index* is computed from the above three measured quantities, S, T, and D. Eq. (23a) was derived in Section III-C.

$$Index = T - f(T-S) - D. \tag{23a}$$

Values of $f(T-S)$ are given in Table II as a function of $T-S$. When the difference $T-S$ exceeds 15 db, $f(T-S)$ is zero and may be ignored.

Reasonable precautions commonly associated with sensitive noise measurements should be followed in operating the test set. The operating location should be free of strong magnetic and electrostatic fields and sources of high-energy radiation, such as RF fields, etc. The test set need not be operated in a screened room. The location should be free from strong mechanical vibrations and loud sources of sound. These precautions are mentioned to serve as a guide in selecting suitable locations. In any case, the suitability of a location can be determined by comparing test set performance in the selected location with that obtained in a "quiet" location. Usually, sources of interference are readily identifiable.

It is not necessary for the test set to be left on continuously to maintain stable operation. However, it is recommended that regular measurements be made of the half-power pass band to assure satisfactory performance. When the resulting accumulation of data indicates stable operation, such measurements may be made at less frequent intervals.

Accuracy of the measured *Index* is dependent, of course, upon the proper calibration of the measuring instrument and upon the stability of the instrument. A feasible means of monitoring the stability of operation is to keep a record of the measurements made on a group of "control" resistors. It is desirable for the control group to consist of different types of resistors and represent a large range of resistance values. Plotting the data in the form of a control chart is suggested as a simple and effective means for displaying irregularities.

VII. Conclusion

The measuring system described in this paper represents the ultimate objective of an NBS resistor-noise study, namely the development of an accurate and reproducible method for evaluating current noise generated in fixed resistors. The program was initiated at NBS eight years ago under sponsorship of the Department of the Navy. The accumulation of experience gained during development of the recommended measuring system includes a study of resistor-noise characteristics, spectral and amplitude characteristics and statistical behavior of measurements of sample groups of resistors; construction and performance testing of several test sets; evaluation of noisiness of about 10,000 resistors representing composition, deposited carbon, metal film and metal-oxide film types.

The measuring system provides a rapid and convenient method for obtaining meaningful resistor-noise information. The information is meaningful because the measured *Index* is a definite quantity, and for this reason is of value for computing current noise in applications.

The *Index* is expressed in decibels because it tends to have a normal distribution when sample groups of "identical" resistors are measured. The resulting collection of data is useful in characterizing the population of resistors from which the sample group has been drawn, in terms of the mean and standard deviation of the measured values of the *Index*.

The measuring system described has proven to be a satisfactory method for the evaluation of resistor noise and is recommended for adoption as a standard test method for the evaluation of current noise generated in fixed resistors.

VIII. Appendix

The current-noise *Index* may be used to calculate the magnitude of current-noise voltage to be expected in specific applications. The equations for such calculations, derived in Section II, will be used in an example to demonstrate the procedure for calculating resistor noise from knowledge of the current-noise *Index*.

The application to be considered is an audio-type amplifier having a half-power pass band from 20 cps to 20 kc, *i.e.*, a three-decade pass band. The problem is to determine the noise voltage to be expected from the plate, cathode, and grid resistors in the first stage of the amplifier. The first stage consists of a low-noise vacuum tube operated in a common-cathode configuration and having an unbypassed cathode resistor. The collective effect of the total noise generated by the three resistors will then be added to an assumed value of tube noise to obtain the total amplifier noise represented as an equivalent noise voltage generator located in the grid circuit of the input tube.

The noise generated by the cathode resistor and the plate and grid resistors will be calculated on the basis of a series of assumptions representative of a given group of typical components. For the input tube, the value of μ is 40, the plate current, 3 ma, and the grid leakage current, 10 nanoamperes. The tube noise is assumed to be equivalent to 2 μv referred to the grid. Pertinent data for the three resistors are summarized in Table IV.

The noise generated by each resistor is considered separately below.

TABLE IV

ASSUMED DATA FOR SAMPLE COMPUTATIONS

Resistor	Resistance Ω	*Index* db	V_{dc} volts
plate, R_L	20 K	0	60
cathode, R_k	1.3K	−10	4
grid, R_g	100. K	0	10^{-3}

A. Noise Generated by the Plate Resistor

The rms voltage value of current noise is given by (7c) in Section II.

$$\sqrt{\overline{v^2}} = V\sqrt{n}\ 10^{(Index)/20} \tag{7c}$$

V is the dc voltage and n is the number of frequency decades of the pass band. For this example, n equals 3. Substituting into (7c) the numbers given in the table for R_L, we obtain the current noise voltage

$$\sqrt{\overline{v^2}} = 60\sqrt{3}\ 10^0 = 104\ \mu\text{v}. \tag{26}$$

The thermal noise produced at the same time by R_L may be computed using Nyquist's equation:

$$\sqrt{\overline{v_0^2}} = \sqrt{4kTR\Delta f} \times 10^6 \tag{27}$$

where k is Boltzmann's constant, 1.38×10^{-23} joules per degree; T is temperature in degrees Kelvin (300°K for room temperature); R is resistance of the resistor in ohms, $20K$; and Δf is the pass band in cps, which equals 2×10^4 cps. The factor 10^6 converts the unit from volts to microvolts. The thermal noise voltage is

$$\sqrt{\overline{v_0^2}} = 2.5\ \mu\text{v}. \tag{28}$$

The total rms noise voltage produced by R_L is the root-sum-square (rss) of the independent noise sources, thermal noise and current noise. Since the current-noise voltage is approximately 40 times that of thermal noise, the thermal noise may be neglected. The total noise produced by R_L may be represented by a noise-free resistor having an equivalent zero-impedance noise voltage generator connected in series with it. The rms noise voltage would be equal to the current-noise voltage, 104 μv.

Before the total effect of noise from all sources may be considered, it is necessary to refer the location of equivalent noise generators to a common location, usually the grid. For noise generated by the plate resistor, this may be achieved by dividing by μ. Hence, the equivalent generator voltage, referred to the grid, is 104/40, or 2.6 μv.

B. Noise Generated by the Cathode Resistor

Substituting into (7c) the numbers given in Table IV for R_k, we obtain the current-noise voltage:

$$\sqrt{\overline{v^2}} = 4\sqrt{3}\ 10^{-1/2} = 2.2\ \mu\text{v}. \tag{29}$$

The thermal noise voltage, obtained from (27), is

$$\sqrt{\overline{v_0^2}} = 0.66\ \mu\text{v}. \tag{30}$$

The current-noise rms voltage is approximately 3.3 times that of thermal noise. The total rms noise voltage of R_k is the rss of the two values,

$$\sqrt{\overline{v^2}}_{(\text{Total})} = \sqrt{(2.2)^2 + (0.66)^2} = 2.3\ \mu\text{v}. \tag{31}$$

This shows that the total noise voltage consists mainly of current noise.

Since the equivalent noise generator of the cathode resistor produces a fluctuation voltage between grid and cathode, its effect, for our purposes, is the same as being located in the grid circuit.[14]

C. Noise Generated by the Grid Resistor

Again, by substituting in (7c) the numbers given in the table for R_g, we obtain the current-noise voltage:

$$\sqrt{\overline{v^2}} = 10^{-3}\sqrt{3}\ 10^0 = 0.0017\ \mu\text{v or } 1.7 \text{ nanovolts}. \tag{32}$$

For this calculation, the dc voltage appearing across the grid resistor is estimated to be 10^{-3} volts. This value is very much smaller than the 150-volt value recommended for measuring the *Index*. Therefore, the measured value of the *Index* should be recognized as being an approximation when applied to the application in question. Since the ratio of the dc voltage of the application to the dc voltage used in measuring the *Index* is approximately 10^{-5}, Fig. 1 may be used to provide an *Index* correction factor associated with a change of 10^{-5} volts. Taking α to be a commonly observed value of 1.9, we obtain an expected *Index* change of 5 db. This would be added to the measured *Index* to obtain the estimated *Index* value for the dc applied voltage equal to 10^{-3} volts. The corrected value of the noise calculated in (32) would then be 5 db larger, *i.e.*, 3.0 nanovolt. For the case in question, the correction is unnecessary, even if an extreme value of α were chosen, such as 1.6, because the resulting correction of a 20-db increase would still indicate that the current noise is negligible.

The thermal noise voltage is

$$\sqrt{\overline{v_0^2}} = 5.7\ \mu\text{v}. \tag{33}$$

For the grid resistor, unlike the other two resistors, thermal noise is the only significant noise source.

[14] The effective grid voltage is actually larger by the additional amount $1/\mu$ owing to the additive effects resulting from the cathode resistor noise generator being located in *both* grid and plate circuits. This correction is considered negligible and therefore is not included for our purposes.

D. Combined Effect of all Noise Sources

The total effect of the three resistors may now be be expressed in terms of an equivalent noise voltage generator located in series with the grid. The rms voltage of the equivalent generator is equal to the rss of the equivalent noise voltage generators of the three resistors.

$$\begin{aligned}\sqrt{\overline{v^2}}_{\text{(Resistors)}} &= \sqrt{(2.6)^2 + (2.3)^2 + (5.7)^2} \\ &= \sqrt{(3.5)^2 + (5.7)^2} \\ &= 6.7\ \mu\text{v}.\end{aligned} \tag{34}$$

The component caused by current noise is 3.5 μv, and the component caused by thermal noise from the grid resistor is 5.7 μv.

The total noise of the input circuit is obtained by including the noise developed by the tube, 2 μv, in the rss equation:

$$\sqrt{\overline{v^2}}_{\text{(Total Input)}} = \sqrt{(3.5)^2 + (5.7)^2 + (2)^2} = 7.0\ \mu\text{v}. \tag{35}$$

If the resistors used were free of current noise, the 3.5-μv term would be zero, and resulting total noise would be 6.0 μv, *i.e.*, a reduction of approximately 1.4 db.

These results represent operation of the amplifier with a signal source having a resistance large compared to the grid resistance, 100$K\Omega$. Consider the other extreme where the driving resistance is small (say less than 4$K\Omega$) compared to the grid resistance. Then the noise component (5.7 μv) which represents thermal noise for the 100$K\Omega$ resistor is reduced to that value associated with a resistance equal to the parallel combination of the grid resistor and driving source resistance, a resistance value less than 4$K\Omega$. As a consequence the resulting thermal noise voltage ($< 1.1\ \mu$v) is negligible.

$$\sqrt{\overline{v^2}}_{\text{(Total Input)}} = \sqrt{(3.5)^2 + (2)^2} = 4.0\ \mu\text{v}. \tag{36}$$

In this case, if the resistors were free of current noise and the term 3.5 μv were zero, the resulting noise voltage would decrease from 4.0 to 2 μv, a 6-db decrease.

The results obtained for the above example should not be considered typical of audio-type amplifiers. The significant factors which contribute to the results vary widely. It is important to recognize the significance of the pass band and the fact that the effects of current noise become exceedingly large as the pass band is extended to lower frequencies. For these reasons each application must be considered separately.

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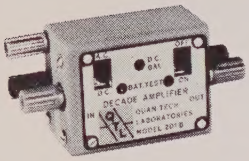
QUAN-TECH INSTRUMENTS

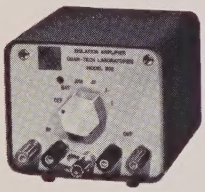


Model 201-B DC-Coupled Decade Amplifier

Highly compact and fully self-contained, the Model 201-B is a stable ac and dc amplifier providing a controlled gain of 10 over a frequency range of dc to 1 MC (±0.5db). Built in battery-condition test indicator. Two or more Model 201-B's may be plugged together in series for greater gain.

Price: $95.00

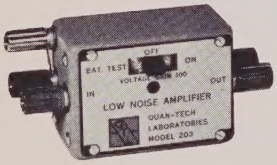
ADAPTORS: Type 1001, Banana Plug to Banana Plug, Price $2.00; Type 1002, Banana Plug to UHF Plug, Price $4.50; Type 1003, Banana Plug to BNC Plug, Price $4.50.



Model 202 Isolation Amplifier

Provides dc and capacitive isolation between input and output terminal pairs. When used with voltmeter, oscilloscopes etc., permits accurate, safe, off-ground measurements. Frequency response: 30 cps to 100 kc (±0.5db). Voltage gain, 1, 0.1, 0.01, 0.001. Battery operated; fully self-contained. Built-in battery condition indicator. Input-to-output effective capacitance, less than 4 mmf. In-phase signal rejection 5000/1 at 100 kc and 10k ohm generator.

Price: $155.00



Model 203 Low-Noise Amplifier

Compact, very low-noise ac amplifier with controlled gain of 100. Frequency response flat from 30 cps to 100 kc (±0.5db); down 3 db at 500 kc. Equiv. noise, full bandwidth, 5 μv RMS, input shorted. Narrow band noise approx. 0.005 μv/cycle above 1 kc. Battery operated; fully self-contained; built in battery condition indicator.

Accessory Adaptor plugs:—See Model 201-B above.

Price: $125.00

Model 317 Amplitude Distribution Analyzer —

Establishes the amplitude-probability distribution of a random signal; for example, to determine threshold requirements, false alarm probabilities, etc. Frequency range: 5 cps to 0.5 mc; Amplitude ranges: 100%, 10% and 1% full scale accuracy. Output to drive counter at front panel. Size 8½"(h) x 5" (w) x 6¾"(d).

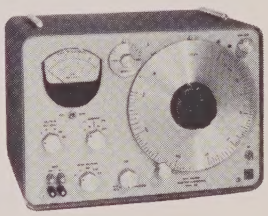
Price: $395.00

Model 301 AC Microammeter

Highly versatile ac current meter providing low input impedance, very high sensitivity, and isolation from ground for low-level current measurements without disturbing circuit under test. Signal output permits display of current waveforms on oscilloscope. Ideal for transistor or magnetic circuit analysis. Sensitivity: With insertion probe, 3 ua to 100 ma full scale; with clamp-on probe, 300 ua to 100 ma, full scale. With 10-amp clamp-on probe (supplied separately) 300 ua to 10 amps.

Price: $290.00

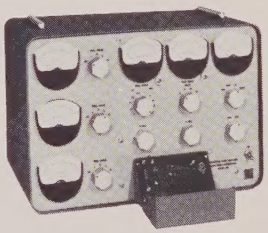
Accessory 10-amp clamp-on probe and connector. **Price: $ 65.00**



Model 303 Wave and Noise Spectrum Analyzer

Designed especially for use in ultrasonics and noise spectrum analysis. Selection of four constant bandwidths; i.e., 10 and 30 cycles (round top), 100 cps and 1 kc (flat top). Frequency range 30 cps to 100 kc, ±300 cps vernier. Sensitivity: 100 μv to 300 v full scale. Analog outputs provided for external monitoring. Sensitivity of 1 μv full scale with Quan-Tech Model 203 amplifier.

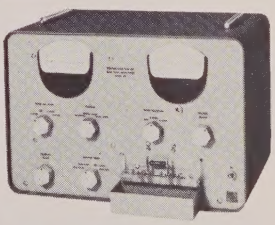
Price: $1425.00



Model 310 Transistor Noise Analyzer

Provides simultaneous analysis of transistor noise at three frequencies. System automatically corrects for variations in transistor gain. Beta, Icbo and Iebo are also included—providing measurements of essential transistor characteristics in one instrument.

Price: $2075.00



Model 315 Resistor Noise Test Set

For precise, quantitative measurement of noise generated as a result of current passing through a resistor. Tests any resistor from 100 ohms to 22 megohms; conforms to recommended specifications of National Bureau of Standards for resistor noise measurements. Tests diodes and other two-terminal networks as well. Outputs provided for external monitoring or for data processing.

Price: $1700.00

Quan-Tech
LABORATORIES, INC.
BOONTON, NEW JERSEY
Phone: (Code 201) DEerfield 4-8500

We welcome requests for information regarding specific applications of Quan-Tech instruments, amplifiers and other products.

Prices and specifications are subject to change without notice. Prices noted are f.o.b. Boonton, N.J., USA.

QUAN-TECH POWER SUPPLIES

Transistorized Power Supplies

A series of compact, lightweight power supplies featuring high regulation and voltage metering. The Model 104B (shown) also includes current metering. Regulation for models 101 and 105, 0.25%; the remaining models, 0.1% regulation. Electronic overload and short-circuit protection.

Model 101	0- 8 VDC, 0-2 amps	Price: $150.00
Model 102	0-14 VDC, 0-1 amp	Price: 150.00
Model 103	0-30 VDC, 0-½ amp	Price: 150.00
Model 104B	0-50 VDC, 0-1 amp	Price: 280.00
Model 105	0-50 VDC, 0-¼ amp	Price: 150.00
Model 108	0-32 VDC, 0-½ amp	Price: will advise

100 SERIES

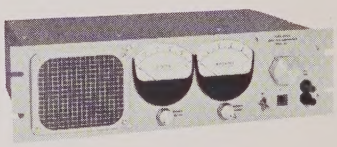
Model 104B

Precision Power Supplies

The 120 Series of precision, solid-state power supplies is characterized by 0.01% regulation. Ripple is less than 500 μvolts, rms. Outstanding features are output voltage and current metering, overload and short circuit protection circuitry, preadjustment of current levels, and remote error sensing.

Model 121	0.1-15 VDC, 0-5 amps	Price: $395.00
Model 122	0.1-36 VDC, 0-3 amps	Price: 395.00
Model 123	0.1-50 VDC, 0-2 amps	Price: 410.00
Model 124	0.1-50 VDC, 0-5 amps	Price: 510.00

120 SERIES



Model 124

Constant Current Supply

Provides constant output current variable over a wide range, independent of voltage across the load. Ideal for transistor or diode testing, potentiometer noise measurements, strain gage applications, etc.

Price: $290.00

OUTPUT: 0.05 to 500 ma. DC in four ranges: 0.05 to 0.5 ma; 0.5 to 5 ma; 5 to 50 ma; and 50 to 500 ma. Maximum output 20 volts.
POWER INPUT: 105-125 volts, 50-60 cycles, 35 watts.

MODEL 151B

Modular Power Supplies

The 170 Series are extremely compact, solid-state power supplies offering excellent operating characteristics *at unusually low cost.* All models are available in 3 configurations: Suffix B for bench use, Suffix C with all connections through octal plug, and Suffix T with terminal strip. All models regulated to less than 3 millivolts, line or load. Ripple is less than 250 μvolts, rms.

Model	Output Voltage	Output Current	Dimensions	Price
170	10-32	0.25a	3″(h)x4″(w)x5″(d)	$ 98.00
171	10-32	0.50a	3.5″(h)x5″(w)x5.5″(d)	123.00
172	0-32	0.20a	3″(h)x4″(w)x5″(d)	105.00
173	0-32	0.40a	3.5″(h)x5″(w)x5.5″(d)	129.00

170 SERIES

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BOONTON, NEW JERSEY
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In addition to the above standard units, Quan-Tech Laboratories produces custom-designed power supplies for specialized applications. We invite the opportunity to offer consultation on your special power supply problems. Address inquiries to the Power Supply Division, at the address below.

The Instrument Division of Quan-Tech Laboratories produces equipment for the measurement and analysis of component and system noise, as well as specialized amplifiers for instrumentation. For details, see section 2900.

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